WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

SEPTEMBER 1956

Vol. 33 No. 9 · THREE SHILLINGS AND SIXPENCE

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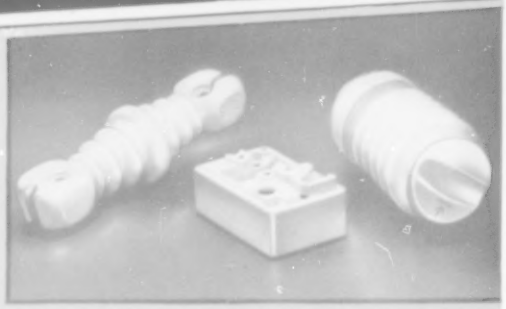

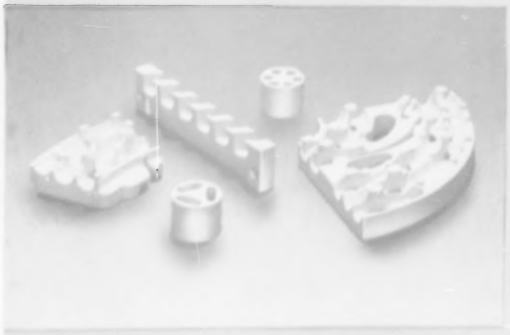

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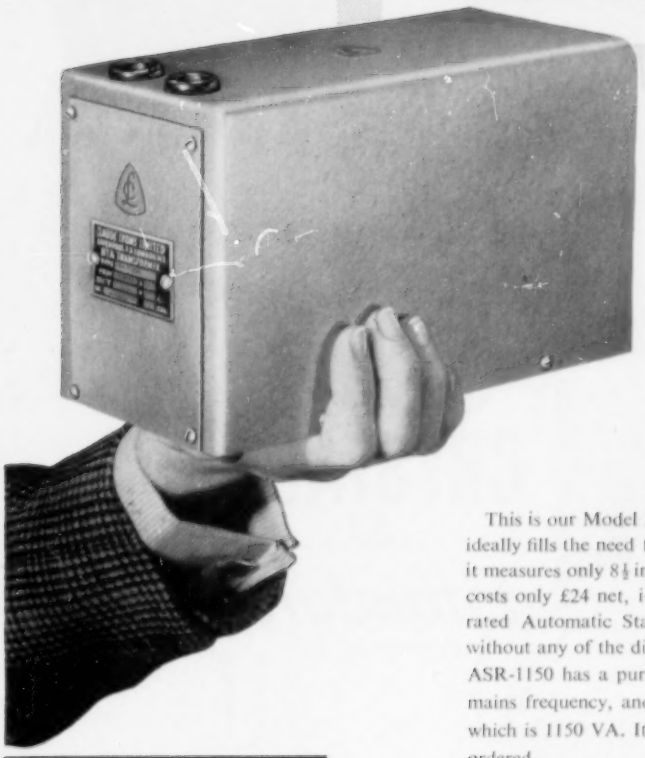

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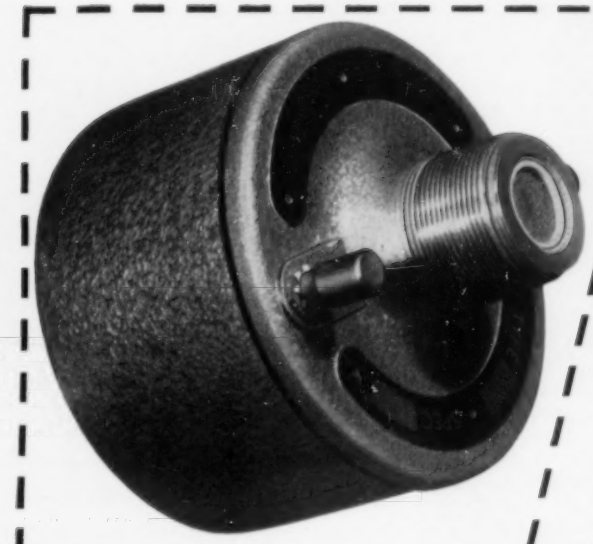

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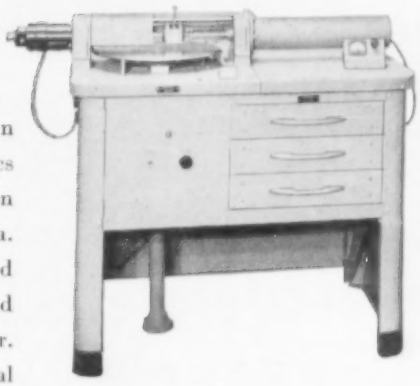

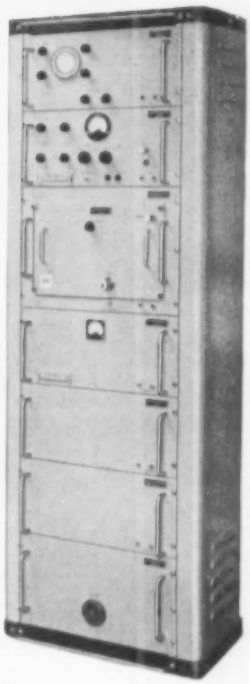

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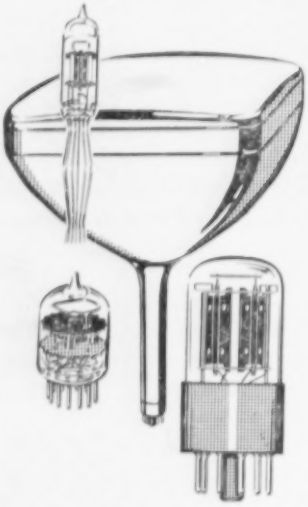

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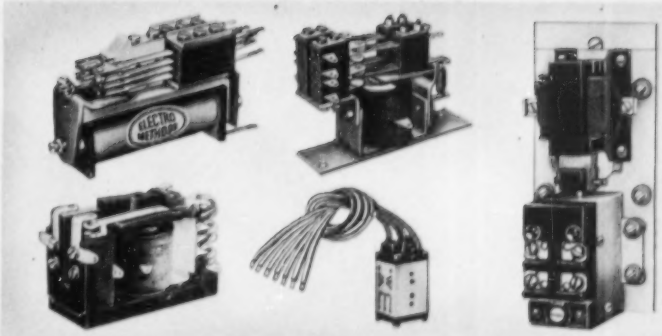

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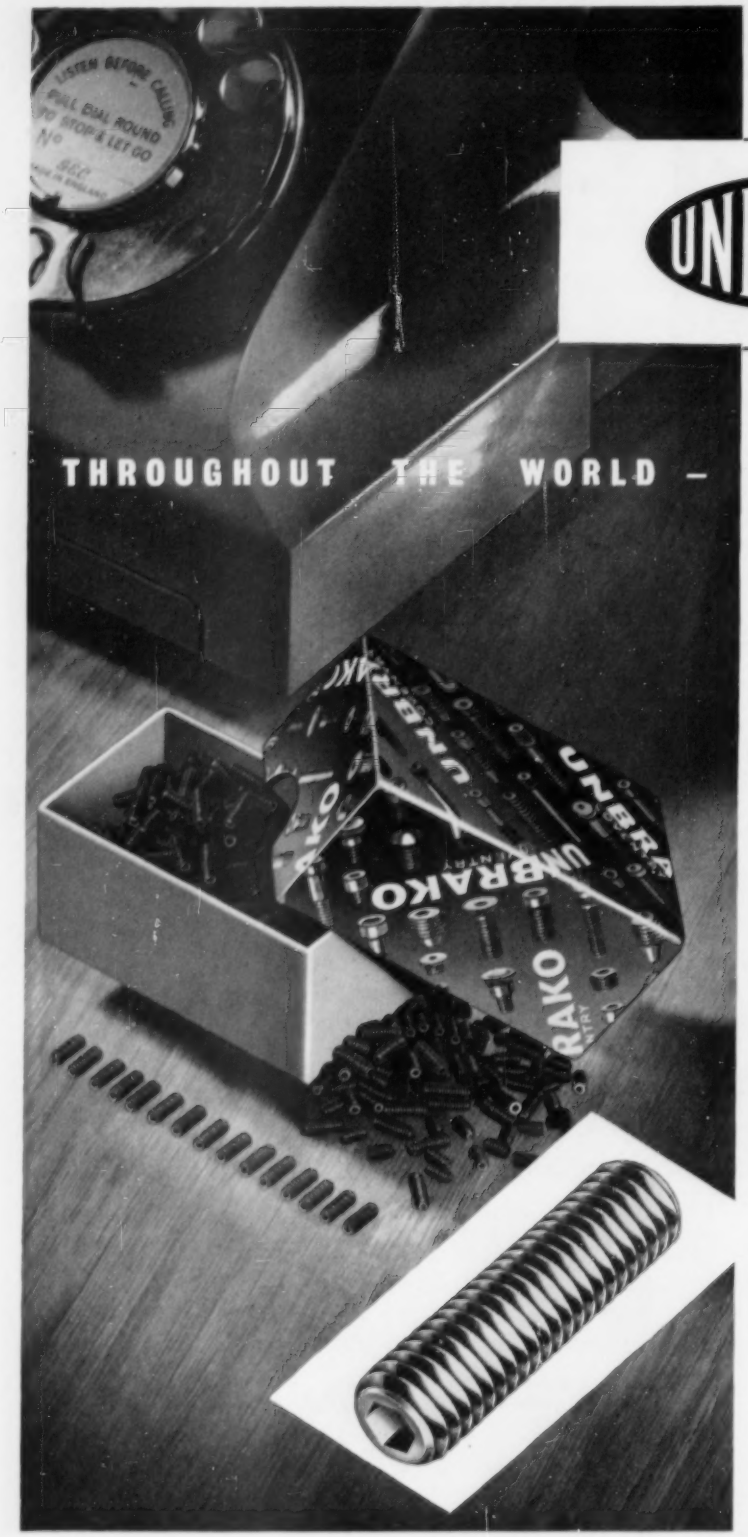

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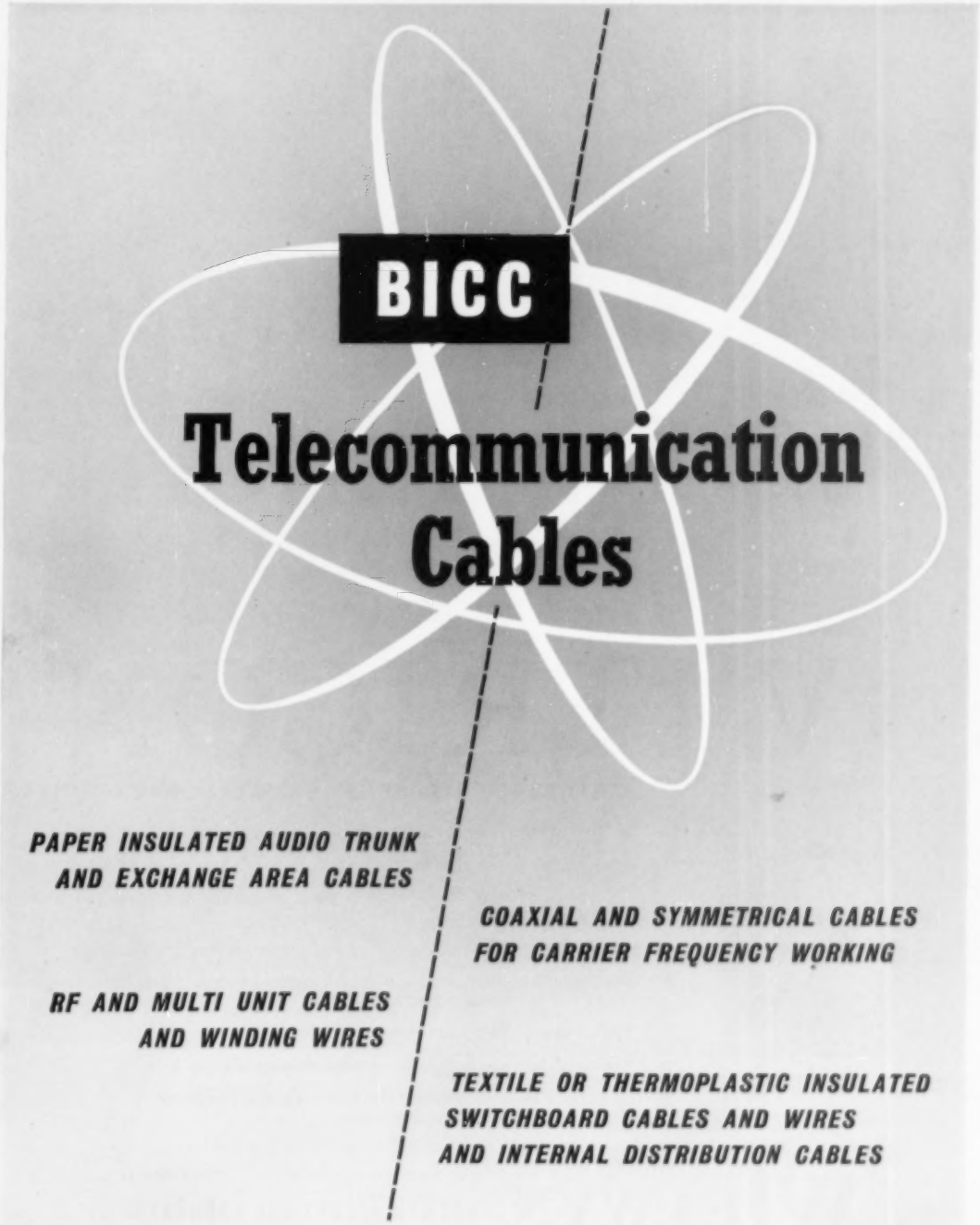
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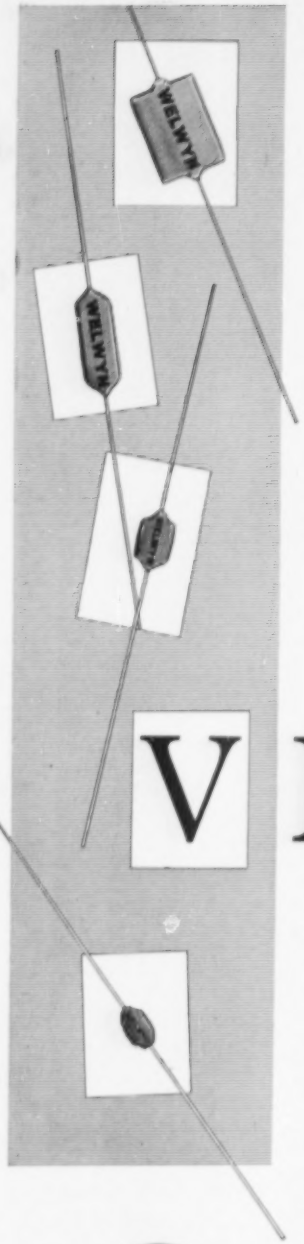

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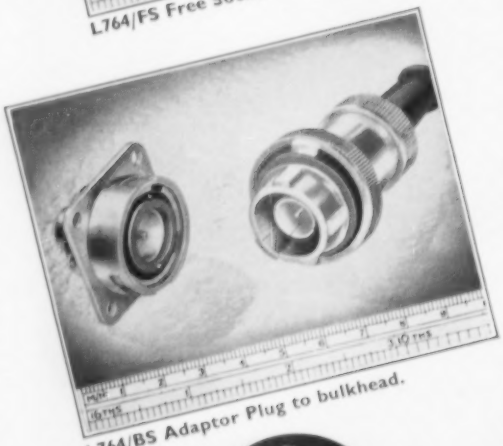

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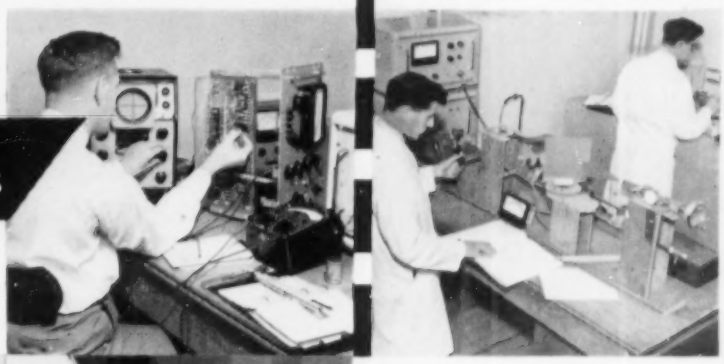

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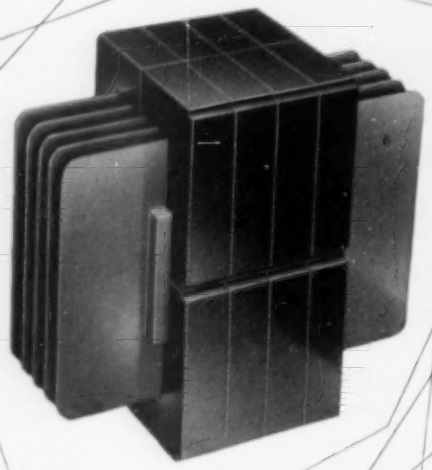

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Vol. 33 SEPTEMBER 1956 No. 9

Janet

KNOWN by the code name of Janet, a new long-range communication system has been developed by the Canadian Defence Research Board. Ranges of up to 1,000 miles with frequencies of 30 – 60 Mc/s are obtained using the trails of minute meteors as the reflecting medium.

The meteors and their ionized trails, which provide the reflections, are natural phenomena and it has long been known that they reflect radio waves. Indeed, in radio-astronomy, radar methods are used for tracking large meteors.

Communication by ionospheric scatter is, of course, well known and is considered to depend in part upon meteor trails. It provides more-or-less continuous communication but demands the use of high power.

Janet employs relatively low power and so permits communication only when the reflecting medium is in just the right state. This is when a meteor trail occurs at the right point in space relative to transmitter and receiver. The period of communication may be as short as a few tenths of a second or as long as a few seconds. However, it has been found that such bursts of communication may occur hundreds of times an hour.

So far as the medium is concerned, communication is highly irregular and the development which has taken place is in apparatus which is capable of utilizing such a variable path for the passing of messages. All messages are recorded on punched paper tape and are stored until communication is established and are then sent at high speed (of the order of 1,300 w.p.m.) while communication lasts. At the receiver, the message is recorded on magnetic tape and typed out in the intervals between communication.

A transmitter and receiver are used at each station and all operate continuously. When a meteor trail occurs in the right region, transmission takes place in both directions and the resulting outputs from the receivers trigger off the modulators and the messages stored at both ends are passed. When transmission is interrupted, the receiver outputs cease and the messages are stopped to await the next suitable meteor trail.

The transmitters, being produced by Redifon Ltd., have a power output of some 800 W and are used with Yagi aerial systems having a gain of about 10 dB.

The idea is certainly an ingenious one and we await further details with interest. At the moment, information is scanty and we feel some hesitation in commenting on it. We are a little disturbed, however, about the use of 40–60 Mc/s for such purposes. This is in the television band, and there would seem to be distinct possibilities of serious interference with television. However, much depends on field-strengths, bandwidths and precise frequency allocations and it may be that there is really no cause for disturbance. At 1,300 w.p.m., the bandwidth needed is clearly quite small and there would seem to be little difficulty in allocating frequencies so that they fall in the guard bands between the television frequencies.

W.T.C.

IMPEDANCE TRANSFORMERS

Experimental Results with Non-Uniform Lines

By J. Willis, B.Sc.(Eng.), A.M.I.E.E. and N. K. Sinha, Ph.D., B.Sc.(Eng.)*

(Faculty of Technology, University of Manchester)

SUMMARY.—The equipment required for the experimental investigation of the use of non-uniform transmission lines as impedance transformers is discussed and a termination developed to give negligible reflection over a wide frequency band. The reflection patterns for two types of tapered line have been obtained experimentally and compared with the theoretical patterns. The effect of attenuation in a practical non-uniform line impedance transformer has been shown to be negligible.

LIST OF SYMBOLS

x = Distance from the sending end of the non-uniform line
l = Length of the non-uniform line
$y = x/l$
$Z_0(x)$ = Nominal characteristic impedance of the non-uniform line
Z_1 = Value of Z_0 for $x = 0$
Z_2 = Value of Z_0 for $x = l$
λ = Free-space wavelength
Γ = Reflection-coefficient
$P(y) = d/dy\,\{\log_e Z_0\}$
d_0 = Inner diameter of the outer conductor of coaxial line
d_c = Outer diameter of the centre conductor of coaxial line
α = Attenuation constant
β = Phase-shift constant
γ = Propagation constant
R = Resistance per unit length of the line
L = Inductance per unit length of the line
C = Shunt capacitance per unit length of the line
G = Shunt conductance per unit length of the line

1. Introduction

IN a paper published recently†, the theory of non-uniform line impedance transformers was discussed and methods were suggested for the design of such lines for optimum performance. There it has been pointed out that an exact solution of the differential equations for non-uniform lines is possible only in a few special cases. This has led to the development of a number of approximate solutions. Among these, the method using Fourier transforms provides the best approximate solution as regards simplicity and accuracy.

It can be shown that the reflection-coefficient at the sending-end of a loss-free non-uniform line impedance transformer, matched at the load-end, is given by

$$\Gamma = \int_0^1 \tfrac{1}{2}P(y)e^{-j4\pi yl/\lambda}\,dy \qquad \ldots \qquad \ldots \qquad (1)$$

It has been shown in the previous paper that optimum distributions of $P(y)$ can be obtained by finding a suitable Fourier series for this function, the coefficients of the various components of the series being adjusted so as to reduce the height of the side-lobes of the reflection pattern† to the desired value.

While the theory of non-uniform line impedance transformers has been discussed in a number of papers, very few give any experimental evidence in confirmation of the theory. Moreover, the theory used involves some approximations. It is, therefore, interesting to investigate the properties of practical non-uniform lines under actual working conditions.

This paper gives some experimental results obtained for the type of non-uniform lines considered in the previous paper.

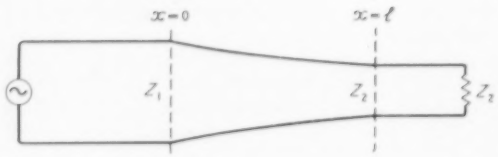

Fig. 1. *Schematic arrangement.*

2. General Description

The most straightforward procedure in verification of the theory is to measure the reflection-coefficient at different frequencies for a given length of taper. The schematic arrangement is shown in Fig. 1.

The non-uniform line is connected between two uniform lines of characteristic impedances equal to those at its respective ends. One of the uniform lines is terminated into its characteristic impedance, Z_2. The measurement of the voltage standing-wave ratio (v.s.w.r.) along the other uniform line at different frequencies will give the reflection pattern for the given non-uniform line.

Practical considerations suggest that it is best to use coaxial lines for the work. The centre conductor of the line can be shaped to obtain any desired taper, keeping the diameter of the outer conductor constant.

* Mr. Sinha is now at the Birla Institute of Technology, Sindri, India.

† "Non-Uniform Transmission Lines as Impedance Transformers", by J. Willis and N. K. Sinha. *Proc. Instn elect. Engrs*, Vol. 103, Part B, No. 8, March 1956, p. 166.

MS accepted by the Editor, November 1955

3. Design of the Measuring Equipment

The main difficulty in the experimental work lies in the inability of the equipment to measure very low standing-wave ratios accurately. Since the main interest is in the side-lobes of the reflection pattern, this limitation becomes very serious. For the best results, therefore, it is essential that the equipment be designed very carefully with the minimum possible inherent reflection. The design of the equipment for the measurement of v.s.w.r. has been discussed in detail in references 2, 5, 10 and 11. By using these techniques, a standing-wave indicator capable of measuring a v.s.w.r. of 1·005 can be obtained.

A further difficulty is the requirement that one of the uniform lines be terminated in its characteristic impedance. The termination is required to give perfect match over the entire range of frequencies at which the measurements are to be made. The two types of termination generally used are (1) the disc resistor and (2) the cylindrical resistor enclosed in a coaxial jacket. The former is easier to design, but it must be backed by a short-circuited quarter-wave line[8] so that the size of the resistor is unduly increased. The latter can be much smaller in size, and does not need any adjustment over a wide range of frequency.

A general study of the cylindrical resistor enclosed in a coaxial jacket of high electrical conductivity, is given in references 6 and 9. In the latter, it has been shown that, making the characteristic impedance of the line formed by the resistor and its jacket about $1/\sqrt{4{\cdot}5}$ times the d.c. resistance of the resistor, and adding a compensating device to cancel the reactance thus caused, it is possible to get a v.s.w.r. of less than 1·01 so long as the length of the resistor does not exceed $\lambda/8$. The most convenient compensating device is a short-circuited transmission line connected in series with the resistor. The schematic arrangement is shown in Fig. 2.

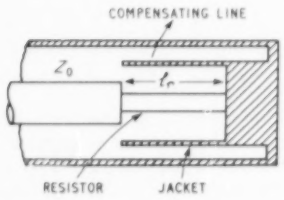

Fig. 2. *Line-compensated resistor.*

By a proper choice of the length of the compensating line and its characteristic impedance[9] a very good impedance match can be obtained over the given range of frequency (i.e., l_r is not greater than $\lambda/8$). Moreover, while designing the compensating line, the discontinuity capacitances caused by the changes in diameters can also be taken into account, thereby cancelling all reactances. Difficulty is experienced, however, in the accurate assessment of the discontinuity capacitances, which are of a very complex nature. The complexity of the discontinuities is further increased by the proximity of the short-circuits in the compensating line and the line formed by the resistor and its jacket. It is also possible that the contribution of the axial-field component in the resistive line to the general field distortion at the junction of the resistive element with the lossless line is not quite negligible. Hence it is not possible to calculate accurately the length of the compensating line that will cancel all reactances. This difficulty can be overcome by making the length of the compensating line adjustable, so that the desired compensation can be obtained under actual working conditions. One adjustment only of the length of the compensating line at the middle frequency in the band will ordinarily be sufficient, but under more rigorous conditions the length of the compensating line can be calibrated to give the best impedance match at all frequencies in the band.

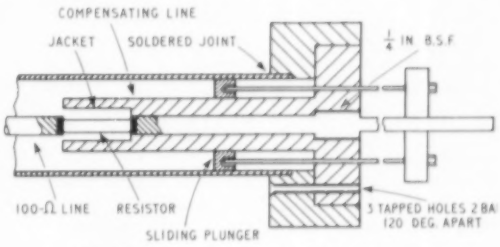

Fig. 3. *Termination for the 100-ohm line.*

The practical adaptation of these principles to terminate a 100-ohm line is shown in Fig. 3. A sliding short-circuiting plunger is used to adjust the length of the compensating line. The adjustment is made by moving the outer disc, the position of which on the central external rod can be calibrated for the frequency at which the best impedance match is obtained.

4. Tests on the V.S.W.R.-Measuring Equipment

To ascertain the accuracy of the equipment, the following tests were performed:

4.1. *Determination of the Standing-Wave Pattern when the Line is Approximately Matched*

The most severe test on the equipment is the determination of the standing-wave pattern when the line is approximately matched at the load-end, since even a very small change in the probe-output alters the standing-wave pattern appreciably. For this purpose, the line was approximately matched using a termination of the type described in Section 3, with inaccurate compensation, and the standing-wave pattern was determined. The standing-wave ratio was found to be

1·015, and even for this low v.s.w.r. the standing-wave pattern was found to differ only slightly from the ideal sine-curve. For example, the standing-wave pattern obtained at the signal frequency of 750 Mc/s is shown in Fig. 4. This indicates that the mechanical irregularities in the equipment are negligible, and that the equipment can be used for the measurement of low standing-wave ratios of the order of 1·01 with sufficient accuracy.

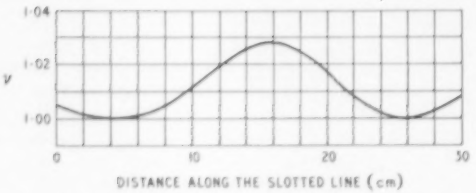

Fig. 4. *Standing-wave pattern at 750 Mc/s when the line is approximately matched; v = voltmeter reading normalized with respect to the minimum value.*

4.2. *Testing the Termination*

The termination shown in Fig. 3 was used to terminate a 100-ohm line and the standing-wave pattern was obtained. The length of the compensating line was adjusted to give the minimum v.s.w.r. in the frequency-band of 450 to 900 Mc/s, and it was found that one setting of the length was sufficient for the entire frequency-band. The reflection-coefficients measured at different frequencies have been plotted against frequency in Fig. 5. It can be concluded that the termination is practically reflection-free in the given frequency-band.

4.3. *Determination of the Standing-Wave Pattern for a Step-Discontinuity in the Centre Conductor, changing the Characteristic Impedance of the Line from 50 ohms to 100 ohms and Terminating the 100-ohm side by its Characteristic Impedance*

This test was performed to obtain an extra check upon the accuracy of the equipment. The schematic arrangement is shown in Fig. 6.

It can be shown theoretically that the standing-wave ratio will be 2·00 in the frequency-band of 450 to 900 Mc/s. The standing-wave patterns actually obtained at these frequencies agreed very closely with the theoretical sine-curves, and the v.s.w.r. was also found to be 2·00.

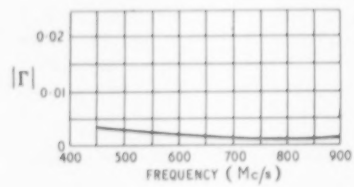

Fig. 5. *Frequency-characteristic of the termination.*

This test, therefore, confirms that the standing-wave detector and the termination give reasonably accurate results.

5. Tests on the Non-Uniform Lines

The experimental determination of the complete reflection pattern for a given non-uniform line involves the measurement of the standing-wave ratio for all frequencies from zero to infinity but, in practice, the measurement can be made only in a limited frequency-range, so that only a portion of the reflection pattern can be determined experimentally. The lower frequency-limit at which the v.s.w.r. can be conveniently measured is determined by the physical length of the slotted line, whereas the upper frequency-limit is determined by its transverse dimensions. In the present work, the frequency-range was limited to the available signal generator range of 450 to 900 Mc/s.

It was decided to verify the theory of non-uniform lines in the case of two types of taper, namely, (1) the simple linear taper of the centre conductor, and (2) the type for which,

$$P(y) \propto (\sin \pi y - \tfrac{1}{6} \sin 3\pi y).$$

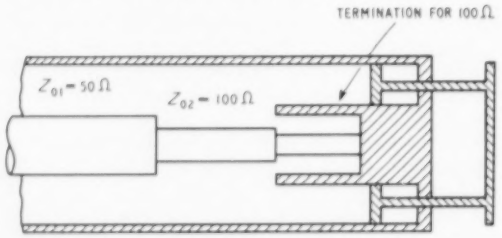

Fig. 6. *Schematic arrangement for the test.*

The former was selected mainly because it is the simplest type of non-uniform line to construct. Another reason for the choice was that, in this case, the side-lobes of the reflection pattern are sufficiently high for the reflection-coefficients to be measured quite accurately. Two lengths of the tapered section were constructed, one 25 cm long and the other 50 cm long, so that a large portion of the reflection pattern could be verified. With the available signal generator, the values of l/λ vary between 0·375 and 1·5 in this case, so that a part of the main-lobe and two side-lobes of the reflection pattern can be obtained experimentally.

The latter was selected mainly because it gives a v.s.w.r. of less than 1·01 for l/λ equal to or greater than one. This line was made 50 cm long, hence the v.s.w.r. will be less than 1·01 in the frequency-band of 600 to 900 Mc/s. A good check on the impedance-matching property of the optimized lines can be obtained thereby.

The combined reflection pattern obtained for the two lengths of straight taper is shown in Fig. 7. It will be seen that the difference between the theoretical and experimental values is very small.

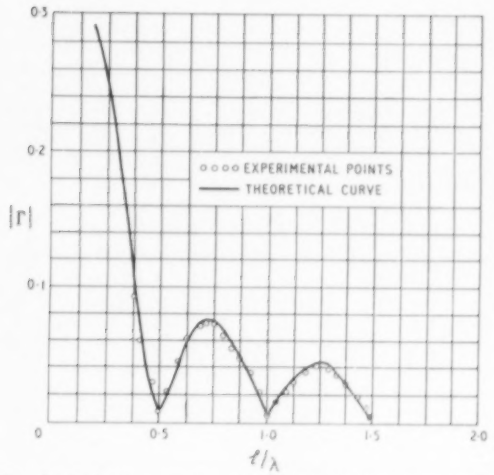

Fig. 7. *Reflection pattern for linear taper of the centre conductor.*

The reflection pattern for the line for which

$$P(y) \propto (\sin \pi y - \tfrac{1}{6} \sin 3\pi y)$$

is shown in Fig. 8. It will again be seen that the experimental points agree very closely with the theoretical curve. It may be noted that the small amount of frequency-sensitivity seen in the theoretical curve is almost completely absent from the experimental curve. The reason is that the inherent reflection present in the slotted line due to structural imperfections affects the accuracy of the measurement of standing-wave ratios less than 1·01.

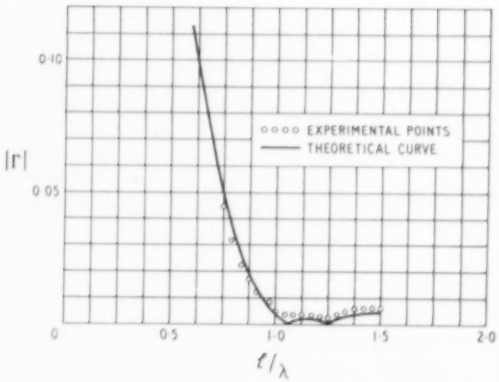

Fig. 8. *Reflection pattern for the line for which* $P(y) \propto (\sin \pi y - \frac{1}{6} \sin 3\pi y)$.

6. Conclusion

In the experimental work carried out, it has been shown that a sudden transition in the characteristic impedance from 50 ohms to 100 ohms causes a standing-wave ratio of 2·0, and that the reflection caused by this change in the characteristic impedance can be reduced if the change is introduced gradually in the form of a non-uniform line of continuously varying parameters.

A linear change in the diameter of the centre conductor to vary the characteristic impedance from 50 ohms to 100 ohms gives a maximum v.s.w.r. of 1·16 for l/λ greater than 0·41 and causes the line to be highly frequency-sensitive. Considerable improvement can be made by tapering the centre conductor to give one of the optimized distributions of $P(y)$. Thus, for l/λ greater than one and

$$P(y) \propto (\sin \pi y - \tfrac{1}{6} \sin 3\pi y),$$

a v.s.w.r. less than the measurable limit of 1·01

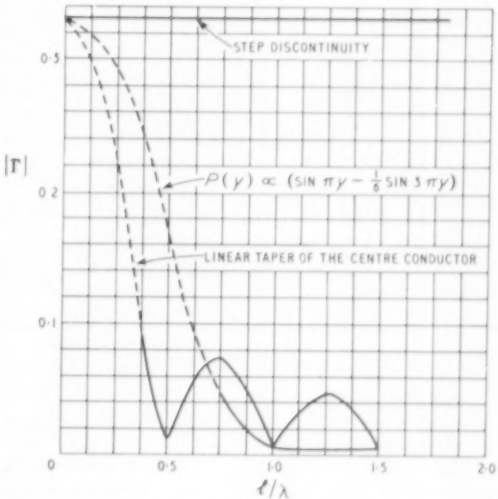

Fig. 9. *Comparison between the various reflection patterns.*

was obtained. These reflection patterns have been shown together in Fig. 9. The portions verified experimentally have been shown solid, and the dotted portions represent the remainder calculated theoretically.

Acknowledgment

The work was carried out in the Electrical Engineering Laboratories of the Faculty of Technology, University of Manchester, and the authors are grateful for the constant interest and encouragement of Professor E. Bradshaw. Thanks are also due to Messrs. A. V. Roe and Company, Limited, of Manchester, for assistance in the

construction of the non-uniform sections of the experimental lines, and to Mr. C. T. Kohn for helpful advice.

APPENDIX 1

Reflection Pattern for Straight Taper of the Centre Conductor

The characteristic impedance of a coaxial line is given by

$$Z_0 = \frac{60}{\sqrt{\epsilon}} \log_e \frac{d_0}{d_c} \quad .. \quad .. \quad .. \quad .. \quad (2)$$

When the diameter of the centre conductor is varied linearly,

$$Z_0 = \frac{60}{\sqrt{\epsilon}} \log_e (a - by) \quad .. \quad .. \quad .. \quad (3)$$

and $$P(y) = -\frac{b}{(a - by) \log_e (a - by)} \quad .. \quad .. \quad (4)$$

where a and b are constants.

Substituting in equation (1),

$$|\Gamma| = \tfrac{1}{2} \int_0^1 \frac{-b}{(a - by) \log_e (a - by)} e^{-j4\pi y l/\lambda} \, dy \quad .. \quad (5)$$

A general solution of equation (5) is not possible, but it can be solved in special cases by finding a suitable approximate expression for $P(y)$. For example, if $Z_1 = 50$ ohms and $Z_2 = 100$ ohms, it can be shown that for an air-dielectric line, with an error of less than 1%,

$$P(y) \approx (0{\cdot}9841 - 0{\cdot}1663\, y + 0{\cdot}2976\, y^2) \log_e \frac{Z_2}{Z_1} \quad .. \quad (6)$$

Using this expression in equation (1), the solution for $|\Gamma|$ was obtained, and its value was calculated for different values of l/λ. The values obtained for integral values of $4l/\lambda$ are given in the following table:

$4l/\lambda$	1	2	3	4	5	6
$\lvert\Gamma\rvert$	0·218	0·0089	0·077	0·0038	0·046	0·0025

APPENDIX 2

Effect of Attenuation

One of the approximations made in deriving equation (1) was that the line was assumed loss free. Although no line can be loss free, all practical high-frequency lines have a low loss. The effect of attenuation in the case of the linear taper of the centre conductor will be investigated here.

If the attenuation in the line is not zero, equation (1) can be re-written as,

$$|\Gamma| = \int_0^1 \tfrac{1}{2} P(y)\, e^{-2\gamma l y} \, dy$$

$$= \int_0^1 \tfrac{1}{2} P(y)\, e^{-2(\alpha + j\beta) y l} \, dy \quad .. \quad .. \quad (7)$$

It can be shown that in a transmission line with low losses, to a very close approximation,

$$\alpha = \frac{R}{2Z_0} + \frac{G}{2} Z_0 \quad .. \quad .. \quad .. \quad .. \quad (8)$$

and $$\beta = \omega \sqrt{LC} = \frac{2\pi}{\lambda} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

For an air-dielectric coaxial line, $G = 0$, and it can be shown that in this case[8]

$$\alpha = \frac{K\sqrt{f}}{d_0} \frac{1 + e^{z_0/60}}{Z_0} \text{ nepers per unit length} \quad (10)$$

where f = frequency in cycles per second
and K = constant depending upon the resistivity of the conductors
$= 83 \times 10^{-9}$ for brass (approximately).

It is evident from equation (10) that for any given value of d_0 the attenuation constant will vary with Z_0. However, the variation is quite small between $Z_0 = 50$ ohms and $Z_0 = 100$ ohms, as shown in the following table:

Z_0 (ohms)	50	60	75	90	100
$\frac{1 + e^{z_0/60}}{Z_0}$	0·0660	0·0620	0·0599	0·0620	0·0629

Hence an approximate solution of equation (7) can be obtained by assuming that the attenuation is independent of Z_0, and the error arising will be small if Z_0 varies between 50 ohms and 100 ohms. For a tapered line 50 cm long, and having d_0 equal to one inch, it will be assumed that α is constant at its value for $Z_0 = 50$ ohms. With this assumption, for brass conductors,

$$\alpha = \frac{0{\cdot}03737}{\sqrt{\lambda}} \text{ nepers per metre} \quad .. \quad .. \quad (11)$$

Substituting the values of α and β in equation (7), it is found that

$$|\Gamma| = \int_0^1 \tfrac{1}{2} P(y) \exp. \left[-2\left(\frac{0{\cdot}03737 l}{\sqrt{\lambda}} + j2\pi l/\lambda \right) y \right] dy$$

$$= \int_0^1 \tfrac{1}{2} P(y) \exp. \left[-(0{\cdot}05285\sqrt{l/\lambda} + j4\pi l/\lambda) y \right] dy \quad (12)$$

taking $l = 0{\cdot}5$ metre.

Substituting the approximate expression for $P(y)$ from equation (6), the values of $|\Gamma|$ can be obtained for different values of l/λ. It was found that attenuation causes a negligible change in the reflection pattern. The values of $|\Gamma|$ calculated with and without attenuation are given in the following table:

$4l/\lambda$	1	2	3
$\lvert\Gamma\rvert$ with attenuation	0·217	0·0090	0·077
$\lvert\Gamma\rvert$ without attenuation	0·218	0·0089	0·077

$4l/\lambda$	4	5	6
$\lvert\Gamma\rvert$ with attenuation	0·0039	0·046	0·0026
$\lvert\Gamma\rvert$ without attenuation	0·0038	0·046	0·0025

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OSCILLATOR FREQUENCY STABILITY

Load Coupled through Long Feeder

By A. S. Gladwin, D.Sc., Ph.D.

(University of Sheffield)

SUMMARY.—It is shown that in addition to the well-known hysteresis effects which occur in an oscillator connected to a mismatched load through a long feeder, another type of instability is also possible. This takes the form of a periodic modulation of the oscillation frequency and can occur when the load is a resonant circuit. A basic theory is developed from which stability criteria appropriate to the two types of instability and to various circuit arrangements are derived. Instability can always be inhibited by choosing suitable values for the circuit parameters. It is shown that, in general, the hysteresis type of instability is likely to predominate.

1. Introduction

IT is well known that frequency instability may arise in an oscillator circuit where a mismatched resistive load is connected to the generator through a long feeder. The instability is manifest in the form of a hysteresis effect when an attempt is made to tune the oscillator. Sudden and irreversible frequency changes take place at certain critical points, and there is a band or bands of frequencies over which the oscillator cannot be tuned at all. This type of behaviour has been described by several writers[1–6] and is fairly well understood.

However, a recent study[7] has disclosed, as a general possibility, the existence of another type of frequency instability. This takes the form of a periodic frequency modulation and may be initiated by adjusting some parameter of the circuit towards a critical value. This has been called 'periodic instability' and, unlike the 'aperiodic' type of instability responsible for hysteresis, periodic instability is reversible with respect to changes in the parameters whose adjustment produces it.

It is natural to enquire whether this new type of instability can exist in oscillators with long feeders, and the answer is that it may arise under certain conditions when the feeder is terminated by a resonant load. Thus, the criteria for the absence of hysteresis are not by themselves sufficient to guarantee stability; further conditions must be imposed if the frequency is not to be periodically unstable.

Problems of stability also arise when an oscillator is coupled through a comparatively short feeder to a resonant cavity (e.g., for the purpose of stabilizing the mean frequency) or to other devices in which the reactance changes rapidly with frequency. It is the object of this paper to determine the conditions necessary for complete stability. A theory of the 2-terminal oscillator is first worked out in general terms and the results are applied to the circuit arrangements involving feeders which arise most frequently in practice.

MS accepted by the Editor, October 1955

2. List of Symbols

$F(p)$	= Frequency stability function. [see (3.12)]
l	= Length of feeder.
n	= Number of quarter-wavelengths in feeder.
p	= $\alpha + j\omega_d$. Characteristic complex frequency of the transient disturbance.
Q_0, Q_1	= External Q-factor of oscillator coil, Q-factor of load coil.
Q_0', Q_1'	= Modified values of Q_0 and Q_1. [see (4.12)]
R	= Impedance of oscillator network at frequency ω_0.
R_0	= $1/Y_0$. Characteristic impedance of feeder.
R_1	= Load resistance referred to feeder.
r_v	= Ratio of group to phase velocities in the feeder.
v_g	= Group velocity of waves in the feeder.
z	= pl/v_g.
Z^+, Z^-	= Impedance of the oscillator network at the complex frequencies $p + j\omega_0$, $p - j\omega_0$.
γ	= $\alpha + j\beta$. Propagation constant of the feeder at frequency ω_0.
γ^+	= Propagation constant of feeder at frequency $p + j\omega_0$.
λ	= Wavelength of a wave of frequency ω_0 in the feeder.
ω_0	= Frequency of steady oscillation.

3. General Theory

The procedure adopted is to calculate first the possible steady states of the oscillator; i.e., the states for which the system is in equilibrium with a periodic wave of constant amplitude and frequency. A small disturbance is then added to the steady state. If the magnitude of the disturbance decreases with time the equilibrium is stable and the possible steady state may be an actual steady state. If the disturbance tends to increase with time, equilibrium is unstable and the possible steady state cannot be physically realized.

Fig. 1 shows a 2-terminal oscillator consisting

Fig. 1. *2-terminal oscillator.*

of a linear passive network Z connected across a nonlinear resistance element NL.R. which has a negative slope over part of its current-voltage characteristic. The current i in the element is assumed to be a continuous and single-valued

function of the voltage v. Thus $i = f(v)$. Z is assumed to have a significant value only near to the oscillation frequency so that all harmonic, low-frequency and direct voltages can be neglected.

For a steady oscillation the voltage across Z is therefore $v = V \cos \omega_0 t$. In the nonlinear element the current of fundamental frequency is $i = I \cos \omega_0 t$ where

$$I = (2/\pi)\int_0^\pi f(V \cos x)\cos x \, dx = Vh(V) \quad (3.1)$$

This must be equal and opposite to the current in Z. Hence if $Z = R + jX$, the equation of the possible steady states is $-V/(R + jX) = Vh(V)$, or

$$X = 0 \text{ and } Rh(V) = -1 \quad .. \quad .. \quad (3.2)$$

the function $h(V)$ being defined by (3.1).

Suppose now that a small disturbance v_d is added to the steady-state voltage v. The current in the nonlinear element becomes $f(v + v_d) = f(v) + v_d f'(v)$. Since $f(v)$ is a periodic function of $\omega_0 t$, $f'(v)$ must also be periodic in $\omega_0 t$. Thus

$$f'(v) = \sum_{-\infty}^{\infty} G_n \exp(jn\omega_0 t) \quad .. \quad .. \quad (3.3)$$

where $G_n = (1/\pi)\int_0^\pi f'(V \cos x) \cos nx \, dx \quad (3.4)$

From (3.1), (3.2) and (3.4) it can be shown that

$$G_0 - G_2 = -1/R \quad .. \quad .. \quad .. \quad (3.5)$$

Using the differential operator D $(= d/dt)$, the current in the network Z due to v_d can be written $[1/Z(D)]v_d$, and this is equal and opposite to the current in the nonlinear element also due to v_d. Hence

$$[1/Z(D)]v_d = -v_d f'(v) \quad .. \quad .. \quad (3.6)$$

If the linear network has lumped parameters $Z(D)$ is a rational function of D — the quotient of two polynomials. In this case (3.6) is a linear differential equation with periodic coefficients the solutions of which are known[8] to be of the form

$$v_d = \exp(at) \sum_{-\infty}^{\infty} V_n \cos[(n\omega_0 + \omega_d)t + \theta_n]$$

or, in complex notation,

$$v_d = V_d \exp(pt) \sum_{-\infty}^{\infty} u_n \exp(jn\omega_0 t) \quad .. \quad (3.7)$$

The actual voltage is to be taken as the real part of the last expression and $p = a + j\omega_d$ is a complex frequency characteristic of each solution. In general, the complete solution of (3.6) would be the sum of a number of such expressions with different values of p and u_n.

When the network has distributed parameters as, for example, when the load is coupled through a uniform feeder, the situation is less simple. A straightforward method would be to write down the partial differential equations for the feeder and to apply the boundary conditions appropriate to the two ends. However, a fresh analysis is unnecessary if only for the reason that the physical behaviour of a network with distributed parameters can always be approximated, to any required degree of accuracy, by a network with lumped parameters. It is perhaps also helpful to regard a transient state as a kind of steady state in which the system is in equilibrium with a group of sine waves having exponential amplitude factors. For all such waves the impedance of a linear network, whether with lumped or distributed parameters, is a constant quantity dependent only on the complex frequency of the applied wave. If the network contains distributed parameters, $Z(D)$ contains hyperbolic functions and (3.6) is not a differential equation in the ordinary sense. Nevertheless, the foregoing remarks indicate that the solutions are still of the form (3.7).

The values of p which satisfy (3.6) and (3.7) are found as follows. Since the network impedance is assumed to be negligible, except near to the oscillation frequency, only those components of v_d which are near to ω_0 need be considered.

$$v_d = V_d[u_1 \exp(p + j\omega_0)t + u_{-1} \exp(p - j\omega_0)t] \quad .. \quad .. \quad (3.8)$$

In order that no terms associated with harmonic frequencies should fall within the frequency band represented by this expression the imaginary part of p must be restricted so that the frequency ranges $p \pm \omega_0$, $p \pm 2\omega_0$ etc., do not overlap. Thus $-\frac{1}{2}\omega_0 < \omega_d < \frac{1}{2}\omega_0$.

Now $[1/Z(D)]\exp(p + j\omega_0)t = \exp(p + j\omega_0)t/Z(p + j\omega_0)$.

Hence the left-hand side of equation (3.6) becomes $V_d[(u_1/Z^+)\exp(p + j\omega_0)t + (u_{-1}/Z^-)\exp(p - j\omega_0)t]$ where $Z^+ = Z(p + j\omega_0)$ and $Z^- = Z(p - j\omega_0)$. The terms of corresponding frequencies on the right-hand side of (3.6) are, from (3.3) and (3.8) $-V_d(u_1 G_0 + u_{-1}G_2)\exp(p + j\omega_0)t - V_d(u_{-1}G_0 + u_1 G_2)\exp(p - j\omega_0)t$.

Equating terms of like frequencies gives

$$\left.\begin{aligned} a_1 u_1 + c_1 u_{-1} &= 0 \\ a_{-1}u_1 + c_{-1}u_{-1} &= 0 \end{aligned}\right\} \quad .. \quad .. \quad (3.9)$$

where

$$\left.\begin{aligned} a_1 &= 1 + G_0 Z^+ & c_1 &= G_2 Z^+ \\ a_{-1} &= G_2 Z^- & c_{-1} &= 1 + G_0 Z^- \end{aligned}\right\} \quad (3.10)$$

In order that solutions of (3.9) other than $u_1 = u_{-1} = 0$ should be possible, it is necessary that $a_1 c_{-1} = c_1 a_{-1}$. Since all the coefficients are functions of p this equation can be written

$$D(p) = a_1 c_{-1} - c_1 a_{-1} = 0 \quad .. \quad .. \quad (3.11)$$

This is the characteristic equation of the transient states whose roots are the complex characteristic frequencies.

When $p = 0$, $Z^+ = Z^- = R$, and by using (3.5) it is easily shown that $D(0) = 0$. Hence $p = 0$ is a solution of (3.11), but this represents a wave of constant amplitude and frequency ω_0 and so corresponds merely to permanent infinitesimal changes in the amplitude and phase of the steady-state oscillation. All other roots represent transients of increasing or decreasing amplitudes. The stability criterion is, therefore, that all the roots of (3.11), other than 0, should have negative real parts, since this ensures that the transients ultimately vanish.

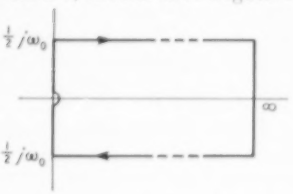

Fig. 2. *p-plane contour.*

Of special interest are oscillators in which the impedance Z is symmetrical with respect to the oscillation frequency; i.e., the real part of Z has even symmetry and the imaginary part odd symmetry with respect to ω_0. This, of course, can be realized only approximately and over a limited range of frequencies. Then $Zj(\omega_0 + \omega_d) = Z^*j(\omega_0 - \omega_d)$, or in terms of complex frequencies, $Z(p + j\omega_0) = Z^*(p^* + j\omega_0)$, where the asterisk denotes the complex conjugate value. Since for all impedances $Z^*(p^* + j\omega_0) = Z(p - j\omega_0)$, it follows that for symmetrical impedances $Z(p + j\omega_0) = Z(p - j\omega_0)$, or $Z^+ = Z^-$.

Assuming symmetry and using (3.5), expression (3.11) can be factorized to give

$$D(p) = F(p)\,A(p)$$

where $F(p) = 1 - Z^+/R$ . . . . . (3.12)

and $A(p) = 1 + (G_0 + G_2)Z^+$

The oscillation is therefore unstable if any root of either $F(p) = 0$ or $A(p) = 0$ has a positive real part. To see the physical significance of these two conditions, suppose that in expression (3.8) $u_{-1} = -u_1$ and let $u_1 V_d = \frac{1}{2}bV \exp(j\theta)$, then the sum of the real part of the transient voltage v_d and the steady-state voltage $V \cos \omega_0 t$ is a frequency-modulated wave

$$V \cos[\omega_0 t + b \exp(at)\sin(\omega_d t + \theta)] \quad . . \quad (3.13)$$

Equations (3.9) reduce to the single equation $F(p) = 0$. Hence, if any root of this equation has a positive real part, the oscillation frequency is unstable.

Similarly, if $u_{-1} = u_1$ the sum of the real part of v_d and the steady-state oscillation is an amplitude-modulated wave

$$V[1 + b \exp(at)\cos(\omega_d t + \theta)]\cos \omega_0 t$$

Equations (3.9) reduce to the single equation $A(p) = 0$, and if this has any root with a positive real part the oscillation amplitude is unstable. As previously explained, the root $p = 0$, which is a root of $F(p) = 0$, corresponds to a small change in the amplitude and phase of the steady-state oscillation. Thus the general transient in a symmetrical-circuit oscillator takes the form of independent modulations of amplitude and frequency. Only frequency modulation is considered further.

Except for the simplest circuits the frequency stability equation $F(p) = 0$ is not readily soluble, but the information required about the nature of its roots can be obtained directly by plotting the locus of $F(p)$ in the complex plane when p moves once round the contour shown in Fig. 2. The frequency is unstable or stable as the locus does or does not enclose the origin. This follows from a theorem of Cauchy according to which the number of times which the locus of $F(p)$ encircles the origin when p describes any closed contour is $N - P$, where N is the number of zeroes and P the number of poles of $F(p)$ lying within the contour of p. Since Z is a passive impedance function, $F(p)$ can have no poles in the right-hand half-plane. Hence each encirclement indicates a root of the equation $F(p) = 0$ with a real positive part. The imaginary part of p is restricted to the range $\pm \frac{1}{2}\omega_0$, for the reason previously discussed, and the root $p = 0$, which is of no interest, is excluded by indenting the contour at the origin as shown in Fig. 2.

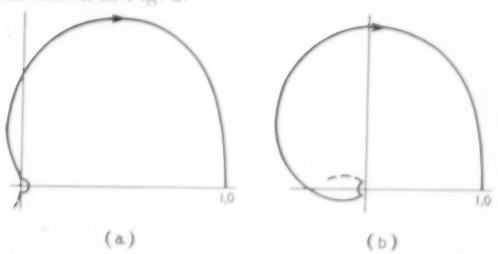

Fig. 3. *Aperiodic stability and instability:* ($n = 1$, $\alpha = 0$). *For (a)* $Q_0 = 60$, $Q_1 = 40$; *and for (b)* $Q_0 = 40$, $Q_1 = 60$.

Typical loci of $F(p)$ are shown in Figs. 3 and 4. Owing to the shunting effect of stray capacitances, Z^+ is small for large values of p, and so $F(p) \to 1$ as $p \to \infty$. Only one half of each locus is plotted, namely from $p = 0$ to $p = \infty + \frac{1}{2}j\omega_0$, the other half being simply a mirror image with respect to the real axis. If the derivative $F'(0)$ [$= dF(p)/dp$ at $p = 0$] is positive, the indentation in the p-plane contour is mapped into the right-hand half-plane as shown in Figs. 3(a), 4(a) and (b). If, however, $F'(0)$ is negative, the indentation falls into the left-hand half-plane as shown in Fig. 3(b). Inspection of these diagrams shows

that the frequency will certainly be unstable if $F'(0) < 0$, for the locus must then make at least one encirclement of the origin. Hence a necessary criterion for stability is

$$F'(0) > 0 \quad .. \quad .. \quad .. \quad .. \quad (3.14)$$

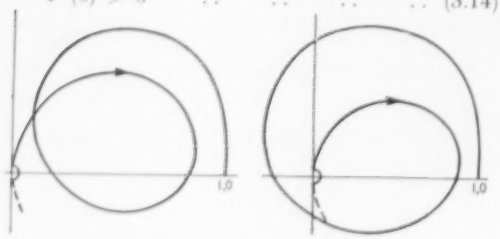

Fig. 4. *Periodic stability and instability;* $(n = 2, \alpha = 0)$. *For (a)* $Q_0 = 10$, $Q_1 = 9$; *(b)* $Q_0 = 9$, $Q_1 = 10$.

In the simplest case of instability corresponding to Fig. 3(b) the equation $F(p) = 0$ has one real positive root $p = a$ and the instability takes the form of a unidirectional change of the frequency away from the value ω_0. This is shown by expression (3.13) when $\omega_d = 0$. Here the instability is of the aperiodic type. Since $F(p)$ is an analytic function of p its differential coefficient with respect to p is the same as with respect to $j\omega_d$. The real part of Z has even symmetry about ω_0 and so at this point $dR/d\omega_d = 0$. Hence, and from (3.12), the stability criterion (3.14) can be written as $-X' > 0$, where $X' = dX/d\omega_d$ at $\omega_d = 0$. This is the familar criterion used by all previous writers, though no very satisfactory proof seems hitherto to have been given. An equivalent form is $-d\phi/d\omega_d > 0$ where ϕ is the phase angle of Z. Although (3.14) has been derived on the assumption of circuit symmetry it is not difficult to show that it is valid also for asymmetrical networks.

It is clear, however, that condition (3.14), although necessary, is not sufficient to ensure stability for, even when the condition is satisfied, the frequency may still be unstable as shown by Fig. 4(b). Here the locus makes two encirclements showing that the frequency-stability equation has two roots (in this case complex conjugates) with positive real parts. This corresponds to periodic instability since the transient now has the form of a periodic frequency modulation of exponentially increasing magnitude as shown by expression (3.13). No very simple criterion for periodic stability can be given but a few special cases are treated analytically in later Sections.

4. Application to the 'Long-Line' Oscillator

For the most part, oscillators in which the load is coupled through a long feeder use magnetron, klystron or other types of microwave generator. It is characteristic of such devices that the equivalent admittance, at the oscillation frequency, of the space charge or electron beam has a susceptive as well as a conductive component, both of which may depend on the amplitude and frequency of oscillation. Thus the internal operation of the generator cannot be represented in any simple way.

A discussion of the effects of electronic admittance variation on frequency stability is beyond the scope of the present inquiry. The purpose of the foregoing remarks is simply to indicate that some error must be expected when stability criteria based on the assumption that the valve can be represented as a nonlinear resistor are applied to magnetron, klystron and similar types of generators.

A further difficulty is that the analysis of the previous Section applies only to continuous-wave steady-state operation, but many microwave oscillators are pulse modulated and this introduces new features into the stability conditions. If the pulse length is shorter than twice the time required to traverse the feeder the load can have no influence on the performance of the generator. When the pulse length is longer than this value the reflected wave (assuming the feeder to be mismatched) reacts on the generator to modify the initial frequency. From Domb's study[9] of this effect is would appear that shortly after the arrival of the first reflection the oscillator has settled down to its steady-state frequency to which the stability criteria may therefore be applied. With these reservations the results of the general theory will be applied to the problem of the long-line oscillator. Only frequency-symmetrical networks are covered by this theory but this is not a serious drawback as will be explained later.

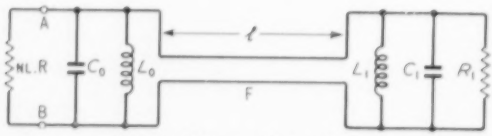

Fig. 5. *Basic oscillator circuit.*

Fig. 5 shows the arrangement to be examined. The generator is represented by the nonlinear resistance element NL.R. and the resonant circuit L_0C_0. Any dissipation in this circuit can be incorporated into NL.R. The network $L_1C_1R_1$ represents the load which is coupled to the generator through a uniform feeder F of length l. In practice the feeder would be connected at both ends through transformers or impedance-matching networks, but the effect of these is simply to alter the scale of the load and

generator impedances and to introduce new reactive elements which, as a first approximation, can be incorporated into the load and generator impedances and into the feeder length. The impedances shown in Fig. 5 are therefore these transformed or 'referred' impedances. Although Fig. 5 is by no means a general or exact representation of the actual oscillator it is sufficiently accurate at frequencies near to ω_0, and only such frequencies are of interest for the present purpose.

The impedance Z of the general theory is the impedance measured between terminals A and B with NL.R. removed. In order that this impedance should be symmetrical with respect to ω_0 the two resonant circuits must be tuned to ω_0 and the feeder must be an integral number of quarter-wavelengths long at this frequency. Thus

$$L_0C_0 = L_1C_1 = 1/\omega_0^2 \text{ and } l = n\lambda/4 \quad .. \quad (4.1)$$

where λ is the wavelength in the feeder. It might seem that this restriction on the feeder length constitutes a serious limitation, but in most applications the length will seldom be known exactly and, with a long feeder, it may easily vary by several wavelengths when the oscillation frequency is varied. It is necessary and sufficient to consider the extreme conditions and these arise when the feeder is an odd or even number of wavelengths long.

At the complex frequency $p + j\omega_0$ the admittance of the load circuit is

$$Y_1^+ = 1/R_1 + 1/(p + j\omega_0)L_1 + (p + j\omega_0)C_1$$

Assuming $|p|$ to be small compared with ω_0, the admittance can be written as

$$\left.\begin{aligned} Y_1^+ &= (1 + 2Q_1p/\omega_0)/R_1 \\ \text{where} \quad Q_1 &= R_1/\omega_0L_1 \end{aligned}\right\} \quad .. \quad (4.2)$$

Similarly the admittance of L_0 and C_0 is

$$\left.\begin{aligned} Y_c^+ &= 2Q_0p/\omega_0R_0 \\ \text{where} \quad Q_0 &= R_0/\omega_0L_0 \end{aligned}\right\} \quad .. \quad .. \quad (4.3)$$

In these expressions R_0, which is arbitrary, will be taken as the characteristic impedance of the feeder. The value of Q_0 thus defined is sometimes called the 'external Q-factor' since it indicates the extent to which the oscillatory circuit is damped by external loading.

Two parameters suffice to specify the steady-state performance of the feeder—the characteristic admittance Y_0 and the propagation constant $\gamma = \alpha + j\beta$. For coaxial or 2-wire lines at high frequencies Y_0 is resistive ($= 1/R_0$) and substantially independent of frequency. At the complex frequency $p + j\omega_0$ the value of the propagation constant is obtained by replacing $j\omega_0$ by $p + j\omega_0$ in the expression for γ. Thus

$$\gamma^+ = [\{R + (p + j\omega_0)L\}\{G + (p + j\omega_0)C\}]^{\frac{1}{2}}$$

where R, G, L and C have the usual significance (line constants per unit length). Ignoring the slight variation of R and G with frequency and assuming, as is usual, that $R \ll \omega_0L$ and $G \ll \omega_0C$ this expression becomes

$$\gamma^+ = \gamma + p\beta/\omega_0 = \gamma + p/v_p$$

where v_p is the phase velocity of waves in the line.

This result can easily be generalized to cover any uniform feeder, for if p is small

$$\gamma^+ = \gamma + p(\partial\gamma/\partial p) = \gamma + p/\partial(\alpha + j\beta)/\partial j\omega_d]$$

In practical feeders α is small compared with β and its rate of change with frequency is altogether negligible compared with the rate of change of β. Hence $\gamma^+ = \gamma + p(\partial\beta/\partial\omega_d)$; but the quantity $\partial\beta/\partial\omega_d$ defines the group velocity v_g, and so

$$\gamma^+ = \gamma + p/v_g \quad .. \quad .. \quad .. \quad (4.4)$$

In the case of the coaxial line the phase and group velocities are equal under the assumed conditions. Hence (4.4) is valid in all cases.

For all practical waveguides α is small and v_g at the frequency ω_0 is given by

$$v_g = v(1 - \omega_c^2/\omega_0^2)^{\frac{1}{2}} \quad .. \quad .. \quad .. \quad (4.5)$$

where v is the phase velocity in the unbounded medium and ω_c the cut-off frequency of the guide. The phase velocity of waves in the guide is $v_p = v(1 - \omega_c^2/\omega_0^2)^{-\frac{1}{2}}$ and the ratio of the two velocities is

$$r_v = v_g/v_p = 1 - \omega_c^2/\omega_0^2 \quad .. \quad .. \quad (4.6)$$

From (4.1) $l/v_g = n\pi/2r_v\omega_0 \quad .. \quad .. \quad (4.7)$

For waveguides the characteristic admittance Y_0 must be replaced by the wave admittance Y_w defined in terms of the transverse electric and magnetic forces. Unfortunately Y_w varies with frequency, but except near to the cut-off frequency the variation is not rapid. Thus for all H-waves

$$Y_w^+ = Y_w[1 - jp/\omega_0(\omega_0^2/\omega_c^2 - 1)]$$

ω_c will usually be of the order of $0{\cdot}6\omega_0$, and since $|p| \ll \omega_0$ the variation of Y_w^+ with p can be neglected.

If a uniform feeder of length l with the parameters Y_0 and γ^+ is terminated by an admittance Y_1^+, the sending-end admittance is

$$Y_s^+ = Y_0(Y_1^+ + Y_0 \tanh \gamma^+l)/(Y_0 + Y_1^+ \tanh \gamma^+l) \quad .. \quad .. \quad (4.8)$$

The total admittance of the network to the right of terminals A,B in Fig. 5 is $Y_c^+ + Y_s^+ = 1/Z^+$. Using (4.2), (4.3), (4.4) and (4.8) the frequency stability function $F(p)$ can be written as

$$F(p) = 1 - Z^+/R = \frac{a_1z + a_2z^2 + (a_3 + a_4z + a_5z^2)\tanh z}{b_0[b_1 + b_2z + b_3z^2 + (b_4 + b_5z + b_6z^2)\tanh z]} \quad .. \quad .. \quad (4.9)$$

in which $z = pl/v_g = pn\pi/2r_v\omega_0 \quad .. \quad .. \quad (4.10)$

and

$$a_1 = Q_0'(R_1/R_0 + T)^2 + (Q_1'R_1/R_0)(1 - T^2)$$
$$a_2 = Q_0'Q_1'(R_1/R_0 + T)T$$
$$a_3 = \tfrac{1}{2}(R_1/R_0)^2 - 1\}(1 - T^2)$$
$$a_4 = Q_0'(1 + TR_1/R_0)(R_1/R_0 + T) - Q_1'(1 - T^2)$$
$$a_5 = Q_0'Q_1'(R_1/R_0 + T)$$

$$b_0 = R_1/R_0 + T$$
$$b_1 = 1 + TR_1/R_0$$
$$b_2 = Q_0'(R_1/R_0 + T) + Q_1'$$
$$b_3 = Q_0'Q_1'T$$
$$b_4 = b_0$$
$$b_5 = Q_0'(1 + TR_1/R_0) + Q_1'T$$
$$b_6 = Q_0'Q_1' \quad .. \quad .. \quad .. \quad .. \quad (4.11)$$

where $T = \tanh \gamma l$ and

$$Q_0' = 4Q_0 r_v/n\pi \quad Q_1' = 4Q_1 r_v/n\pi \quad .. \quad (4.12)$$

Since the feeder length is assumed to be an integral number of quarter wavelengths at the oscillation frequency, the total phase shift along the feeder is $\beta l = \frac{1}{2}n\pi$. Then

$$\left.\begin{aligned} T = \tanh \gamma l &= \tanh(\alpha l + \tfrac{1}{2}jn\pi) \\ &= \tanh \alpha l \text{ if } n \text{ is even} \\ &= \coth \alpha l \text{ if } n \text{ is odd} \end{aligned}\right\} \quad (4.13)$$

Stability can always be determined by plotting the locus of $F(p)$, but even in the simplest cases this is a very laborious procedure. The stability criterion is also expressed by requiring that all the roots of $F(p) = 0$, other than 0, should lie in the left-hand half-plane. Since the denominator of (4.9) has no poles except at ∞, and since z is a real positive multiple of p, this is equivalent to demanding that all the roots other than 0 of the equation

$$a_1 z + a_2 z^2 + (a_3 + a_4 z + a_5 z^2)\tanh z = 0 \quad (4.14)$$

should lie in the left-hand half-plane. A complete solution of this problem will not be attempted, for in most cases of practical interest one or more of the coefficients is small enough to be neglected and the simplified equation is easily treated.

5. Aperiodic Stability

In Section 3 is was shown that a necessary condition for stability is $F'(0) > 0$, the frequency being aperiodically unstable when this condition is not fulfilled. Since $F(0) = 0$ the derivative can be written as $F'(0) = \lim_{p \to 0} F(p)/p$. Also the product $b_0 b_1$ in the denominator of (4.9) is positive. Hence the criterion for aperiodic stability can be expressed as

$$a_1 + a_3 > 0 \quad .. \quad .. \quad .. \quad .. \quad (5.1)$$

Substituting according to (4.11) and (4.13) gives, if n is even

$$Q_0' > \frac{[1 - (R_1/R_0)^2 - Q_1'R_1/R_0]\operatorname{sech}^2 \alpha l}{(R_1/R_0 + \tanh \alpha l)^2} \quad (5.2)$$

and if n is odd

$$Q_0' > \frac{[Q_1'R_1/R_0 + (R_1/R_0)^2 - 1]\operatorname{sech}^2 \alpha l}{[1 + (R_1/R_0)\tanh \alpha l]^2} \quad (5.3)$$

When n is even the effect of Q_1 is to improve stability, and when n is odd the stability margin is decreased. In cases where the feeder length is not known exactly both (5.2) and (5.3) must be satisfied. The effect of attenuation is always to improve stability. A number of special instances of (5.2) and (5.3) are now considered in detail.

First, suppose that the load is matched to the feeder at the oscillation frequency; i.e., $R_1 = R_0$. Then (5.2) is always satisfied and (5.3) becomes (using (4.12))

$$Q_0/Q_1 > \exp(-2\alpha l) \quad .. \quad .. \quad .. \quad (5.4)$$

This is satisfied for all values of αl if $Q_0 > Q_1$. Fig. 3(a) and (b) show loci of $F(p)$ corresponding to $Q_0 > Q_1$ (stable) and $Q_0 < Q_1$ (unstable) for the case of a loss-free quarter-wave feeder with $r_v = 1$.

Next, suppose that the load is almost entirely resistive over a wide frequency range or that the feeder is very long. In either case $Q_1'(= 4Q_1 r_v/n\pi)$ is negligible and criteria (5.2) and (5.3) become, if n is even

$$Q_0 > \frac{(\pi l/\lambda r_v)[(R_0/R_1)^2 - 1]}{[\cosh \alpha l + (R_0/R_1)\sinh \alpha l]^2} \quad .. \quad (5.5)$$

and if n is odd

$$Q_0 > \frac{(\pi l/\lambda r_v)[(R_1/R_0)^2 - 1]}{[\cosh \alpha l + (R_1/R_0)\sinh \alpha l]^2} \quad .. \quad (5.6)$$

It is seen that when n is even aperiodic instability can exist only when $R_1 < R_0$, and if n is odd instability exists only when $R_1 > R_0$. If the attenuation were zero it would always be possible to find a feeder length such that no matter how closely the load were matched the frequency would be unstable. The presence of attenuation modifies this conclusion, for the right-hand sides of these inequalities, instead of approaching ∞ as l increases, pass through a maximum and then decrease to zero. Thus, depending on the value of R_1/R_0, there exists a critical value of Q_0 for which the oscillation has aperiodic stability whatever the length of feeder.

To find this critical value it will be supposed that the load is fairly well matched; i.e., $R_1/R_0 = 1 \pm d$, where d is small. The maximum values of the right-hand sides of (5.6) and (5.5) can then be calculated approximately and are found to be

$$Q_m = \pi d/\alpha\lambda e r_v \quad .. \quad .. \quad .. \quad (5.7)$$

where e is the base of Naperian logarithms. Hence if $Q_0 > Q_m$ stability is assured for all lengths of feeder. For example, a rectangular waveguide for operation at 10,000 Mc/s might have $\alpha\lambda = 0{\cdot}0007$ and $r_v = 0{\cdot}57$. With $d = 0{\cdot}1$, corresponding to a standing-wave ratio of 1·1, this would give $Q_m = 290$.

In deriving formula (5.7) it was assumed that Q_1' was negligible compared with $(R_1/R_0)^2 - 1$ in (5.2) and (5.3). Formula (5.7) is therefore valid only if the quantity $2Q_1 r_v/n\pi d$ is much less than 1.

If the feeder length is given and the ratio

R_1/R_0 is regarded as the unknown (Q_1 still being taken as 0) it is again possible to find a value of Q_0 for which aperiodic stability is guaranteed whatever the value of R_1. As R_1/R_0 is increased or decreased away from 1 expressions (5.5) and (5.6) tend asymptotically to the limiting value

$$Q_0 > \pi l/\lambda r_v \sinh^2 \alpha l$$

This is equivalent to equation (36) of Käch[6]. In practice no very great attenuation could be tolerated. For example, with a total attenuation of 3 dB at 10,000 Mc/s and with $\alpha\lambda = 0{\cdot}0007$ and $r_v = 0{\cdot}57$, the required value of Q_0 is 22,000 which is impracticably large.

Formulae (5.5) and (5.6) can also be expressed in terms of the standing-wave ratio r in the feeder. If the attenuation is zero $r = R_1/R_0$ or $r = R_0/R_1$ as R_1/R_0 is greater or less than 1. When attenuation is present the standing-wave ratio has no very precise physical significance, but it may be defined formally at any point in the feeder in terms of the amplitudes of the forward and backward travelling waves. Thus at a distance l from the load the standing-wave ratio is

$$r_0 = (r + \tanh \alpha l)/(1 + r \tanh \alpha l) \quad .. \quad (5.8)$$

where r is defined above. Using this expression (5.5) and (5.6) can be reduced to the single criterion

$$Q_0 > (n\pi/4r_v)(r_0^2 - 1) \quad .. \quad .. \quad (5.9)$$

Allowing for the difference of notation this is the same as formulae (9.31) and (9.32) of Pierce and Shepherd[3] who considered only even values of n and zero attenuation. The expression is also equivalent to equations (33) and (34) of Käch[6].

The effect of a mismatched load on oscillator performance is often specified in terms of the 'pulling-figure'—the amount by which the oscillation frequency varies as a load producing a specified standing-wave ratio is moved along the feeder. Using (4.8) and (5.8) it is easily shown that the total change of susceptance at the sending-end terminals of the feeder produced by moving the load is $\pm \frac{1}{2}(r_0 - 1/r_0)/R_0$. Provided this is small compared with $\omega_0 C$, the total change $\Delta\omega$ in the oscillation frequency is $\Delta\omega/\omega_0 = \frac{1}{2}(r_0 - 1/r_0)/Q_0$. Hence if the pulling is not to exceed a certain specified value, $Q_0 > \frac{1}{2}(r_0 - 1/r_0)\omega_0/\Delta\omega$. Comparing this with the previous formulae it is seen that there is no relation between the pulling figure and the criteria for frequency stability.

6. Periodic Stability

The criteria derived in the last Section, though necessary, are not sufficient to guarantee stability; for the frequency may still be periodically unstable when these conditions are satisfied. To decide the question of periodic stability the roots of the characteristic equation (4.14) must be examined in greater detail. This will now be done for three particular cases.

It is first assumed that the load is purely resistive; i.e., $Q_1 = 0$. (4.11) then shows that the coefficients a_2 and a_5 are zero, and the characteristic equation becomes

$$a_1 z + (a_3 + a_4 z)\tanh z = 0 \quad .. \quad .. \quad (6.1)$$

It can be shown[10] that the conditions for all the roots of this equation, other than 0, to lie in the left-hand half-plane are

$$\left.\begin{aligned} a_1 + a_3 > 0 \text{ if } a_4 > 0 \\ a_1 + a_3 < 0 \quad a_4 < 0 \end{aligned}\right\} \quad .. \quad .. \quad (6.2)$$

Since a_4 is positive the first condition applies and this is simply the criterion (5.1) for aperiodic stability. It follows that periodic instability cannot exist with a purely resistive load.

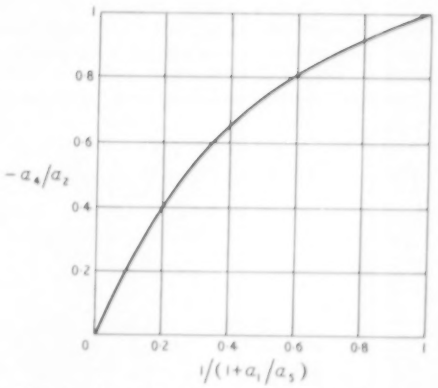

Fig. 6. *Critical values of a_1/a_5 and $-a_4/a_2$.*

Next it is supposed that the load is resonant but matched to the feeder at the oscillation frequency; i.e., $R_1 = R_0$. Then, from (4.11), $a_3 = 0$ and after removing the factor z, the characteristic equation is

$$a_1 + a_2 z + (a_4 + a_5 z)\tanh z = 0 \quad .. \quad (6.3)$$

Assuming that $a_1 > 0$, which is the criterion for aperiodic stability, the conditions for all the roots to lie in the left-hand half-plane are[10]

$$a_2 > 0 \quad a_5 > 0$$

$$\text{and } -a_4/a_2 < y_1 \cot y_1 \quad .. \quad .. \quad .. \quad (6.4)$$

where y_1 is the smallest positive root of the equation

$$y \tan y = a_1/a_5 \quad .. \quad .. \quad .. \quad (6.5)$$

(4.11) shows that a_2 and a_5 are positive and so the first two conditions are always satisfied. The third condition (6.4) can be represented graphically. Fig. 6 shows the critical values of $-a_4/a_2$ calculated from (6.4) and (6.5), plotted as a function of $1/(1 + a_1/a_5)$. Only points lying below the graph represent stable states.

From (4.11) the values of a_4/a_2 and a_1/a_5 with $K_1 = K_0$ are

$$\left.\begin{aligned} a_4/a_2 &= (1 - \coth \gamma l)/Q_0' + (1 + \coth \gamma l)/Q_1' \\ a_1/a_5 &= (1 - \tanh \gamma l)/Q_0' + (1 + \tanh \gamma l)/Q_1' \end{aligned}\right\} \quad .. \quad .. \quad (6.6)$$

If the feeder length is an odd number of quarter-wavelengths (4.13) and the above give

$$a_4/a_2 = (1 - \tanh \varkappa l)/Q_0' + (1 + \tanh \varkappa l)/Q_1'$$

which is always positive. Hence (6.4) is satisfied and periodic instability cannot exist with odd values of n.

If n is even

$$\left.\begin{aligned} a_4/a_2 &= (1 - \coth \varkappa l)/Q_0' + (1 + \coth \varkappa l)/Q_1' \\ a_1/a_5 &= (1 - \tanh \varkappa l)/Q_0' + (1 + \tanh \varkappa l)/Q_1' \end{aligned}\right\} \quad .. \quad .. \quad (6.7)$$

These expressions together with Fig. 6 enable the stability of any given arrangement to be determined but they do not afford a ready means of finding the critical value of any particular parameter; e.g., Q_0. However, with the help of (6.7), expressions (6.4) and (6.5) can be recast in the equivalent forms

$$\left.\begin{aligned} Q_0' &> 2(1 - \tanh \varkappa l)\tan y/y(\tan^2 y + \tanh \varkappa l) \\ Q_1' &< 2(1 + \tanh \varkappa l)\tan y/y(\tan^2 y - \tanh \varkappa l) \end{aligned}\right\} \quad .. \quad .. \quad (6.8)$$

For any given value of the total attenuation $\varkappa l$, corresponding critical values of Q_0' and Q_1' can be calculated by giving y a succession of positive values less than $\frac{1}{2}\pi$ and such that $\tan^2 y > \tanh \varkappa l$. Fig. 7 shows a few graphs obtained in this way. Only points lying above the respective graphs represent stable states as it is only for such points that Q_0' is greater than, or Q_1' is less than, the critical value. This method will give a complete solution of the problem provided only a sufficient number of graphs are constructed to cover the required ranges of $\varkappa l$, Q_0' and Q_1'.

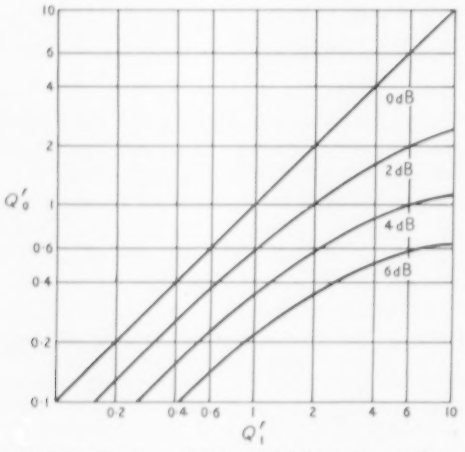

Fig. 7. *Critical values of Q_0' and Q_1' for various values of total attenuation.*

Two general trends can be discerned in Fig. 7. When Q_0' and Q_1' are small the graphs become parallel straight lines with a slope of unity on the logarithmic scale. It follows that the limiting critical values of Q_0 and Q_1 are related by the equation $Q_0/Q_1 = f(\varkappa l)$. When Q_1' is large Q_0' tends asymptotically to a limit which depends only on $\varkappa l$. The value of this limit and of the function $f(\varkappa l)$ will be deduced later. When the attenuation is zero the critical values of Q_0' and Q_1' are equal. Hence if $Q_0 > Q_1$ the frequency has periodic stability for all values of $\varkappa l$. This is also clear from (6.7) which shows that a_4/a_2 is then positive and criterion (6.4) is therefore satisfied. Fig. 4(a) and (b) show loci of $F(p)$ for $Q_0 > Q_1$ (stable) and $Q_0 < Q_1$ (unstable) calculated from (4.9) for a loss-free half-wave line with $r_v = 1$. As shown in the previous Section the condition $Q_0 > Q_1$ also guarantees aperiodic stability for all values of $\varkappa l$ when $K_1 = K_0$.

Periodic stability is also assured for all values of l and Q_1 if

$$Q_0 > \pi/\varkappa\lambda r_v \quad .. \quad .. \quad .. \quad .. \quad (6.9)$$

This can be deduced from expressions (6.8). As y decreases from $\frac{1}{2}\pi$ to its lower limit of $\tan^{-1}(\tanh^{\frac{1}{2}} \varkappa l)$ the critical value of Q_1' increases from 0 to ∞ and the critical value of Q_0' increases continuously from 0 to an upper limit. It follows that stability is assured for all values of Q_1' when Q_0' is greater than this limit. Substituting the lower limit of y into the expression for Q_0' and using (4.12) gives

$$Q_0 > (\pi l/\lambda r_v)(1 - \tanh \varkappa l)/\tanh^{\frac{1}{2}} \varkappa l \tan^{-1}(\tanh^{\frac{1}{2}} \varkappa l) \quad .. \quad .. \quad (6.10)$$

The right-hand side is greatest when $\varkappa l$ is small and this leads at once to (6.9). For a waveguide at 10,000 Mc/s with $\varkappa\lambda = 0{\cdot}0007$ and $r_v = 0{\cdot}57$ the critical value of Q_0 given by (6.9) is 7,900—an impracticably large figure. However, at lower frequencies with coaxial cables more reasonable values of Q_0 are obtained. For example, with the cable used in the experiments at a frequency of 10 Mc/s, the critical value of Q_0 is 39. Although the most severe requirement arises when $\varkappa l$ is small the value of this quantity has no great influence on the critical value of Q_0. Thus for a total attenuation of 3 dB the value of Q_0 for the waveguide is reduced to 5,900.

More definite forms of the relation between the critical values of Q_0 and Q_1 can be obtained for the two limiting cases in which both Q_0' and Q_1' are either large or small. When Q_0' and Q_1' are large a_1/a_5 is small and the solution of equation (6.5) is approximately

$$y_1^2 = (a_1/a_5)(1 - a_1/3a_5) \quad .. \quad .. \quad (6.11)$$

The criterion (6.4) then becomes $-a_4/a_2 < 1 - a_1/3a_5$ which is a good approximation for values of a_1/a_5 up to 0·3. Using (6.7) the criterion can be written

$(3 - 2\tanh \alpha l - \tanh^2 \alpha l)/Q_0' - (3 + 2\tanh \alpha l - \tanh^2 \alpha l)/Q_1' < 3\tanh \alpha l$

If the total attenuation is small $\tanh \alpha l \approx \alpha l$ and the above expression reduces to

$$(1 - 2\alpha l/3)/Q_0 - (1 + 2\alpha l/3)/Q_1 < \alpha\lambda r_v/\pi \quad (6.12)$$

From this the approximate limiting value of Q_0 as $Q_1 \to \infty$ follows at once. The exact value is given by (6.10).

Expression (6.12) may be compared with the criterion for aperiodic stability under the same conditions, namely $R_1 = R_0$ and a small total attenuation. If αl is small the criterion (5.4) can be written $(1 - \alpha l)/Q_0 - (1 + \alpha l)/Q_1 < 0$. If this inequality is satisfied then (6.12) is also satisfied provided $1/Q_0' + 1/Q_1' < 3$, and this is certainly true under the assumed conditions of large Q_0' and Q_1'. It is concluded that in those cases where, because of variations in the oscillator frequency, the electrical length of the feeder may range over an interval of half a wavelength or more, the dominant type of instability under the conditions assumed above is aperiodic.

It is also possible to calculate the characteristic modulation frequency ω_d at the inception of periodic instability. At the critical point between stability and instability the amplitude of the principal transient is stationary, neither increasing nor decreasing, and the characteristic equation has then a pair of purely imaginary roots. Thus $z = \pm jy = \pm \frac{1}{2}jn\pi\omega_d/\omega_0 r_v$. Hence ω_d can be calculated from (6.5) and (6.7). This identification of y with the root of (6.5) owes its validity to the fact that expressions (6.5) and (6.4) (which also becomes an equation in the critical case) are obtained from the characteristic equation (6.1) by writing $z = jy$.

Using the approximate solution (6.11) valid for large values of Q_0' and Q_1'

$$\omega_d/\omega_0 = (2r_v/n\pi)(a_1/a_5)^{\frac{1}{2}}(1 - a_1/6a_5) \quad \ldots (6.13)$$

When a_1/a_5 is not small (6.4) and (6.5) must be solved by the usual methods of successive approximations to find Q_0 and y. A start can be made with the approximate value of Q_0 given by (6.12) or Fig. 7.

Finally, if Q_0 and Q_1 are small or if the feeder length is great then Q_0' and Q_1' are small and the coefficients a_2 and a_5 in the characteristic equation (4.14) are negligible compared with the others. It is no longer necessary to assume that $R_1 = R_0$. The equation becomes

$$a_1 z + (a_3 + a_4 z)\tanh z = 0$$

This is the same as equation (6.1) and the criteria (6.2) therefore apply. Since for aperiodic stability $a_1 + a_3 > 0$ the sole criterion for periodic stability is $a_4 > 0$. From (4.11) and (4.13) it can be seen that if n is odd $\operatorname{sech}^2 \gamma l$ is negative and a_4 is therefore positive. Accordingly, periodic instability is impossible when n is odd. If n is even the stability criterion becomes

$$Q_0/Q_1 > \operatorname{sech}^2 \alpha l/[1 + (R_1/R_0)\tanh \alpha l](R_1/R_0 + \tanh \alpha l) \quad \ldots (6.14)$$

The frequency is stable for all values of αl if $Q_0/Q_1 > R_0/R_1$ or, from (4.2) and (4.3), $C_0 > C_1$. The frequency is stable for all values of R_1/R_0 if $Q_0/Q_1 > 2\operatorname{cosech} 2\alpha l$. If the load is matched to the feeder at the oscillation frequency, $R_1 = R_0$, and (6.14) becomes $Q_0/Q_1 > \exp(-2\alpha l)$. This is the same as the criterion (5.4) for aperiodic stability which applies when n is odd. The physical explanation of this correspondence is that as the feeder length becomes longer and longer the imaginary part $j\omega_d$ of the complex characteristic frequency p becomes smaller and smaller in accordance with (6.13), and so, in the end, periodic and aperiodic instability become indistinguishable.

It seems safe to say that in all cases where the feeder length is not known precisely, stability considerations should be based on the criterion for aperiodic stability, since it is usually aperiodic instability which arises first as Q_0 is decreased.

7. Experimental Results and Conclusions

Although the main application of the theory is probably to microwave oscillators, the apparatus required to test the stability criteria can be more cheaply constructed, and the various parameters more easily and certainly measured, at lower frequencies. Accordingly the experiments were carried out at a frequency of 10 Mc/s. This choice, of course, imposes limitations on the Q-factor values and feeder lengths which can conveniently be used, but there is no reason to suppose that different results would be obtained at higher frequencies.

A triode oscillator was used, as shown in Fig. 8, regeneration being effected by mutual-inductance coupling between the grid and anode coils. Since the theory has been developed only for 2-terminal oscillators, its application to the 4-terminal network of Fig. 8 is perhaps not obvious. However, it can easily be shown that, as far as small frequency changes in symmetrical networks are concerned, the circuit to the left of the dashed line in Fig. 8 (including the mutual inductance) can be replaced by a nonlinear resistor connected across terminals A, B, provided either that the amplification factor of the valve is large or that the coefficient of coupling between grid and anode coils is 1. If the grid-leak and capacitor are replaced by a direct connection or by a grid-bias

battery the equivalence holds also for amplitude changes and asymmetrical networks.

The feeder was a coaxial-cable connected at both ends to taps on the coils. Neglecting losses, a tapped coil behaves as a perfect transformer with an inductance connected in parallel with one pair of terminals and a second (leakage) inductance connected in series with the other pair. Two capacitors were therefore placed in series with the cable as shown to neutralize the leakage inductances at the oscillation frequency and thus to improve the symmetry of the network with respect to frequency. With the values of capacitance used (2,500 – 5,000 pF) the resultant reactances in series with the cable were negligible over a wide frequency range.

By means of a movable dust-core the inductance of the oscillator grid coil, and thus the value of Q_0, could be adjusted, the capacitance being

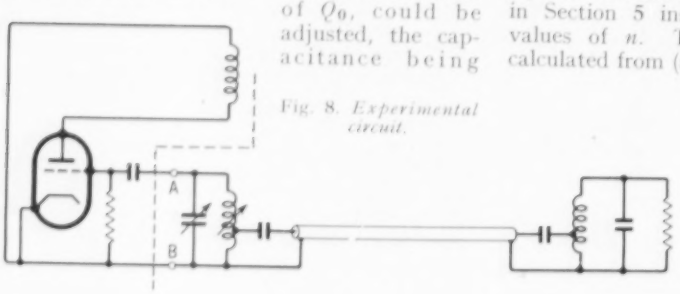

Fig. 8. *Experimental circuit.*

varied simultaneously to keep the circuit tuned to the frequency ω_0. A calibrated receiver with beat-frequency oscillator was used to monitor the oscillation frequency. An extreme roughness of beat note indicated the threshold of aperiodic instability, while periodic instability was manifest in the appearance of two side frequencies corresponding to the modulation frequency ω_d.

In carrying out the experiments the value of Q_0 was slowly decreased until instability was detected. The critical value of Q_0 was then determined by measuring first the intrinsic Q-factor Q_i of the coil with the cable disconnected, and then the total Q-factor Q_L with the cable connected and terminated by its characteristic impedance R_0. From these values the external Q-factor Q_0 is $Q_0 = Q_i Q_L/(Q_i - Q_L)$. The experimental results were as follows.

At a frequency of 10 Mc/s the constants of the cable were $R_0 = 73\,\Omega$, $\alpha = 0{\cdot}00407$ neper/metre, $\lambda = 19{\cdot}6$ m, $r_v = 1$. From these figures, $\alpha l = 0{\cdot}02n$ and $\alpha\lambda = 0{\cdot}08$ neper.

Test 1. Aperiodic stability with resistive load ($Q_1 = 0$). The load circuit (including the series capacitor) shown in Fig. 8 was removed and replaced by a resistor R_1. For two values of R_1 the critical values of Q_0 were found for various lengths of line. Theoretical values of Q_0 were calculated from (5.5) and (5.6).

TEST 1

R_1 (ohms)	n	Q_0	
		Meas.	Calc.
19	2	16·8	16·3
	4	26·1	25·3
	12	35·3	33·8
237	3	16·5	15·7
	5	21·9	21·1
	11	28·4	27·0

Test 2. Aperiodic stability with resonant load ($R_1 = R_0$). For this test the load was matched to the cable at the oscillation frequency. As shown in Section 5 instability exists only with odd values of n. Theoretical values of Q_0 were calculated from (5.4).

TEST 2

n	Q_1	Q_0	
		Meas.	Calc.
3	25·5	23·2	22·6
5	25·5	20·6	20·9
11	38·9	25·3	25·1

Test 3. Periodic stability with resonant load ($R_1 = R_0$). The circuit arrangement was the same as in test 2 but only even values of n were used. At the inception of instability the modulation frequency ω_d was also measured.

TEST 3

n	Q_1	Q_0		$\omega_d/2\pi$ (Mc/s)	
		Meas.	Calc.	Meas.	Calc.
2	38·9	17·7	18·8	1·08	1·10
2	25·5	14·8	14·8	1·20	1·26
4	38·9	17·2	18·0	0·79	0·76
10	38·9	16·2	15·6	0·46	0·46

Theoretical values of Q_0 and ω_d were calculated from (6.12) and (6.13) except for the last result which was obtained by solving (6.4) and (6.5) by successive approximations as described in Section 6.

In carrying out this test a type of hysteresis effect was observed with $n = 2$ and $Q_1 = 25{\cdot}5$. When Q_0 had been decreased to the critical value at which instability set in (and it is these values which are tabulated) it was found that in order to restore stability Q_0 had to be increased to a value a few per cent higher. This effect is associated with the large value of modulation frequency

(1·26 Mc/s) for at this frequency the network impedance–frequency characteristic has become markedly asymmetrical, and frequency modulation is then inevitably accompanied by amplitude modulation. Frequency instability by itself is independent of the properties of the maintaining system, but when amplitude modulation is also present the non-linear characteristic of the amplifier is brought into play. At lower modulation frequencies where the network impedance–frequency characteristic is more highly symmetrical no hysteresis effect could be detected.

It was also found that when Q_0 was reduced well below the value at which instability began a second critical point was reached where the frequency changed suddenly to a new stable value close to one of the side-frequencies $\omega_0 \pm \omega_d$. This was due to the inevitable asymmetry of the network characteristic which favoured one of the side-frequency components at the expense of the other.

In all three tests the theoretical and experimental results are in reasonable agreement. Tests 2 and 3 confirm the conclusion reached in Section 6 that for approximately the same lengths of feeder, e.g., $n = 10$, $n = 11$, the critical value of Q_0 required for aperiodic stability is much greater than for periodic stability.

In practical applications, particularly at microwave frequencies, the results would be influenced by irregularities in the impedance and propagation constant at the feeder and by reflections from junctions and bends. Also, as mentioned previously, the equivalent circuit of the oscillator may not be quite so simple as the basic scheme of Fig. 5. Again it may not always be possible to obtain the desired high value of Q_0. The generator oscillatory circuit L_0C_0, which has justifiably been assumed loss-free for the purpose of the analysis, has in fact an intrinsic Q-factor Q_i. If Q_0 becomes comparable with Q_i much of the total available power is dissipated in the oscillatory circuit instead of being transferred to the load. Hence in practice Q_0 cannot be more than a small fraction of Q_i.

From the practical point of view the problem of a microwave transmitter connected to its radiator by a long waveguide has already been solved in a very satisfactory manner[11] by introducing a unilateral ferrite transducer between the generator and the guide. This allows free transmission of energy to the radiator but suppresses all waves reflected back to the generator, thus making instability impossible. However, for longer wavelengths where this technique cannot be applied or for some of the problems mentioned in the Introduction, the theory of stability presented here may still find a useful application.

Acknowledgment

The work here described was carried out at the Institute for Technical Physics in the Swiss Federal Institute of Technology, Zürich. The author has to thank Professor E. Baumann, Director of the Institute, and Dr. G. Neukomm of the Swiss School Board, for the facilities placed at his disposal.

APPENDIX

Wideband Networks

It was assumed in Section 3 that the network impedance Z has a significant value only near to the frequency ω_0; in particular it was supposed to be negligible along the lines $\alpha \pm \frac{1}{2}j\omega_0$ in Fig. 2. Although this assumption is justified for most practical oscillators it is of some interest to determine what modification is required in the analysis of Section 3 when the network impedance has a significant value over a wide band of frequencies.

Voltages and currents of frequencies p and $p \pm 2j\omega_0$ must now be considered since their magnitudes are comparable with those of frequencies $p \pm j\omega_0$ when p is close to a $\pm \frac{1}{2}j\omega_0$. On the other hand it can be assumed that Z is negligible at frequencies of $2\omega_0$ and beyond. Hence, if attention is confined to positive values of ω_d, it is sufficient to consider four terms in the expression for the complex transient voltage, namely

$$v_d = V_d \exp(pt) \sum_{-2}^{1} u_n \exp(jn\omega_0 t)$$

where, as before, the actual voltage is the real part of this expression.

At the frequency $p + jn\omega_0$ let the network impedance be Z_n and let the current in the non-linear resistor be i_n. Then $i_n = -V_d u_n / Z_n$. But $i_d = v_d f'(v)$ where $f'(v)$ is given by (3.3). Hence the four equations for u_n can be written as

$$\begin{bmatrix} 1+G_0Z_1 & G_1Z_1 & G_2Z_1 & G_3Z_1 \\ G_1Z_0 & 1+G_0Z_0 & G_1Z_0 & G_2Z_0 \\ G_2Z_{-1} & G_1Z_{-1} & 1+G_0Z_{-1} & G_1Z_{-1} \\ G_3Z_{-2} & G_2Z_{-2} & G_1Z_{-2} & 1+G_0Z_{-2} \end{bmatrix} \begin{bmatrix} u_1 \\ u_0 \\ u_{-1} \\ u_{-2} \end{bmatrix} = 0 \qquad \text{A.1}$$

The characteristic frequencies of the transients are given by the equation $D(p) = 0$, where $D(p)$ is the determinant of the first matrix, remembering that only solutions with $0 < \omega_d < \frac{1}{2}\omega_0$ are valid.

If the locus-plot method is used to determine stability only one half of the locus, corresponding to positive values of ω_d, should be drawn: the other half is a mirror image of this. When Z_0 and Z_{-2} are negligible the determinant reduces to (3.11).

It is obvious that Z_0 and Z_{-2} can influence the stability only if Z has a value at the frequencies $\frac{1}{2}\omega_0$ or $1\frac{1}{2}\omega_0$ comparable with its value R at ω_0.

The possibility of the transient wave producing pure frequency modulation may be investigated by writing $u_{-1} = -u_1$ and $u_{-2} = -u_0$. Equations A.1 become

$$\begin{bmatrix} 1+(G_0-G_2)Z_1 & (G_1-G_3)Z_1 \\ 0 & 1+(G_0-G_2)Z_0 \\ 1+(G_0-G_2)Z_{-1} & 0 \\ (G_1-G_3)Z_{-2} & 1+(G_0-G_2)Z_{-2} \end{bmatrix} \begin{bmatrix} u_1 \\ u_0 \end{bmatrix} = 0 \qquad \text{A.2}$$

Since the general solution is any linear combination of the individual solutions of (A.2) it is sufficient to consider these elementary solutions. Apart from the trivial case of $u_1 = u_0 = 0$ three possibilities exist for u_1 and u_0: either may be zero and the other non-zero, or both may be non-zero.

If $u_0 = 0$ then $G_1 - G_3 = 0$, $Z_{-1} = Z_1$ and $1 - Z_1/R = 0$ the last equation following from (3.5). The first condition means that the steady-state current of second-harmonic frequency must be zero; for the amplitude of this current is

$$I_2 = (2/\pi)\int_0^{\pi} f(V\cos x)\cos 2x\,dx$$

while, from (3.4)

$$G_1 - G_3 = (1/v)\int_0^{\pi} f'(V\cos x)(\cos x - \cos 3x)\,dx = 2I_2$$

The second condition requires the network impedance to be symmetrical with respect to ω_0 over the range $\frac{1}{2}\omega_0$ to $1\frac{1}{2}\omega_0$, and the third equation gives the complex frequency of the modulation. Stability can be determined by solving this equation or by plotting the locus of $1 - Z_1/R$.

Similarly if $u_1 = 0$ equations (A.2) become $G_1 - G_3 = 0$, $Z_{-2} = Z_0$ and $1 - Z_0/R = 0$. The second equation requires the network impedance to be symmetrical with respect to ω_0 in the ranges 0 to $\frac{1}{2}\omega_0$ and $1\frac{1}{2}\omega_0$ to $2\omega_0$. The complex modulation frequency is now $j\omega_0 + p^*$ where p (with ω_d positive) is the root of the third equation.

Finally u_0 and u_1 may both be non-zero. Since frequency stability or instability can hardly depend on the amplitude of the steady-state oscillation, $G_1 - G_3 = 0$. Then $Z_0 = Z_1 = Z_{-1} = Z_{-2}$, and the complex modulating frequencies are p and $j\omega_0 + p^*$, where p is the root of the equation $1 - Z_0/R = 0$ with ω_d positive. Thus Z must now be symmetrical not only about ω_0 but also, over small ranges, about the frequencies $\frac{1}{2}\omega_0$ and $1\frac{1}{2}\omega_0$. Such a characteristic is not likely to be obtained except by special design.

It must be concluded that periodic instability in oscillators with networks having significant values of impedance at $\frac{1}{2}\omega_0$ or $1\frac{1}{2}\omega_0$ will always take the form of mixed amplitude and frequency modulation. Even when the network impedance has the required symmetry independent modulations of amplitude and frequency can exist only when the steady-state second-harmonic current is zero.

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HIGH-FREQUENCY ELECTRONIC COUNTER

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(Research Department, British Broadcasting Corporation)

SUMMARY.—A binary counter has been developed which is characterized by a higher counting rate than that hitherto achieved. The article outlines the principles involved and describes a practical arrangement.

1. Introduction

SOME of the television systems[1] now under development include the use of a sub-carrier, the frequency of which lies within the limits of the video spectrum.

The sub-carrier is in turn modulated by signals additional to, and independent of, those required by a normal black-and-white system and, in order to reduce cross-talk between the modulated sub-carrier and the black-and-white signals to the minimum, the sub-carrier frequency is usually chosen to be an odd multiple of half the line-scanning frequency.

In such a television system, therefore, it is necessary to relate both line and frame repetition frequencies to that of the sub-carrier. This may be effected by a chain of frequency dividers and multipliers, as illustrated in Fig. 1, where a sub-carrier, of frequency f_s, is generated in the unit A and is fed to a frequency-dividing and multiplying circuit B.

MS accepted by the Editor, October 1955

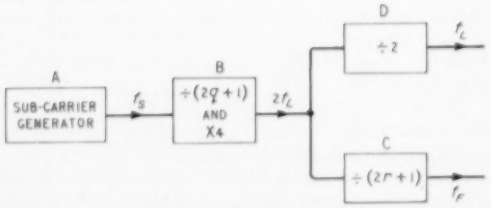

Fig. 1. *Arrangement of counters in a television system employing a band-shared sub-carrier.*

The frequency division effected in B has a divisor which is an odd integer $(2q + 1)$ and the divider output is, in the majority of systems[2], then multiplied by four. With suitable choice of f_s and q, the output from the multiplier can constitute a source of twice line-scan frequency $2f_L$ which may be fed to the customary frame and

line-frequency divider circuits (C and D). In the arrangement illustrated the frame frequency f_F has been related to twice the line frequency by an odd integer $(2r + 1)$, a condition necessary to produce the normal form of interlaced scanning raster.

The sub-carrier frequency chosen for a particular television system may (with British and American scanning standards) lie in the range 2–4 Mc/s, and it is necessary, therefore, that a frequency divider or counter should be developed which will accept an input within such a frequency range and operate satisfactorily when counting by a fairly large odd number which may be prime.

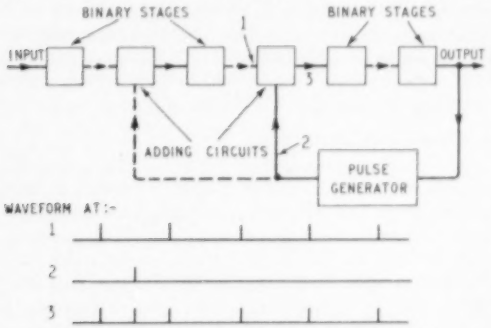

Fig. 2. *Binary cascade with additive feedback.*

2. General Considerations

A well-known form of pulse counter or frequency divider which may be considered to fulfil the requirements listed above is the triggered binary cascade[3].

In such an arrangement, each binary stage delivers one output pulse for every two input pulses. Evidently n such stages arranged in cascade will have an overall divisor or count of 2^n. If pulses derived from the output are reinserted at suitable points, which may include the input, in such a manner that the added pulses occur during the intervals between the normal triggers, the cascade may be arranged to have an overall count equal to any integer between unity and 2^n. Such an arrangement is illustrated in Fig. 2, where the waveform diagrams show the derived feedback pulses added to the normal trigger pulses.

The upper limit to the input pulse repetition frequency, beyond which a binary cascade fails to operate satisfactorily, is determined by two factors. First, due to the presence of stray reactance, triggering of the first binary stage will be followed by a finite recovery period during which the circuit will fail to respond to any further trigger. In these conditions, therefore, the maximum input repetition frequency will be such that the interval between input pulses just exceeds the duration of the first-stage recovery period. Secondly, if feedback pulses derived from the cascade output are added to the input-pulse train, as shown in Fig. 2, the maximum usable input repetition frequency will fall to a value equal to or less than half that acceptable when operating without feedback. The factor of two relating these limits will result when the feedback pulse is arranged to occur exactly midway between two normal trigger pulses.

From the above discussion it will be seen that, in order to satisfy the requirements demanded by the application described in Section 1, the first counter stage must be characterized by a recovery period equal to or less than 0·125 μsec.

A modified form of the binary cascade counter[4] has been developed wherein the second of the two limiting factors described above has been eliminated.

3. Gated Counter

In the modified arrangement a pulse, derived from the output of the cascade, is used to suppress one or more of the normal triggers driving various binary stages in the cascade. A simple example illustrating this principle is shown in Fig. 3, where a cascade of n stages is driven from the output of a gating circuit consisting of a simple electronic switch.

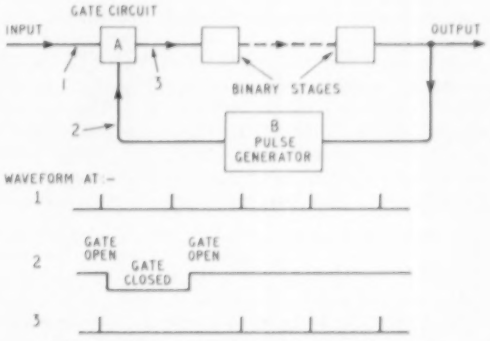

Fig. 3. *Binary cascade with gating feedback.*

Fig. 3 shows an input-pulse train (1) fed to a gate circuit A which is, in turn, operated by a control pulse (2) derived from the cascade output by means of the pulse generator B. The gate output waveform (3) constitutes the drive to a series of binary stages and, if the cascade consists of n such stages, 2^n pulses of the waveform (3) will occur for every output pulse from the cascade. If each control pulse (2) is arranged to remove one pulse from the gate output, then each output pulse from the cascade will correspond to $2^n + 1$ pulses of the waveform (1). These operations are illustrated in the waveform diagrams of Fig. 3.

It follows that if each control pulse is arranged to remove x input pulses, the overall count will be $2^n + x$. If the repetition frequency of the input pulses does not vary by more than one octave, the control pulse (2) may have a constant duration; however, operation may be made independent of input repetition frequency below a certain upper limit if the gate circuit A and the pulse generator B are of the form shown in Fig. 4.

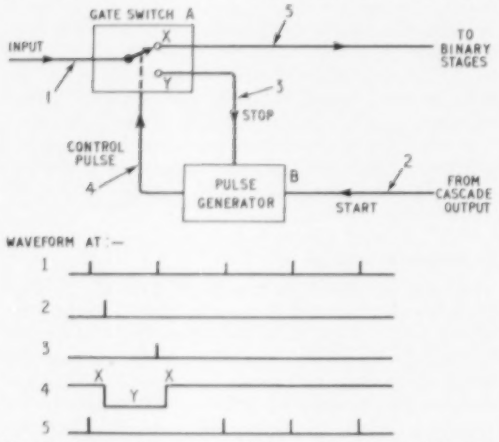

Fig. 4. *Automatic control of the gating pulse duration.*

In this case, the gate circuit A is arranged as a double-throw electronic switch controlled, as before, from the pulse generator B. The switch normally connects the input waveform (1) to the binary cascade (switch position X). The cascade output pulse (2) is fed to the pulse generator B as a 'start' pulse and initiates the control pulse (4) which, in turn, operates the electronic switch (position Y). The next input pulse is thus diverted into the pulse generator as a 'stop' pulse (3). The waveform diagrams of Fig. 4 illustrate these operations and show that the trailing edge of the control waveform always occurs immediately following the trailing edge of the pulse gated out. A pulse counter may be inserted in the 'stop' pulse circuit, in order that the gate circuit may remove two or more pulses per operation.

In a general arrangement several such gate circuits may be inserted at various points in the binary cascade. Each of these gate circuits may then be operated by suitable control pulses derived from the cascade output.

It can be shown (see Appendix) that if such an arrangement contains n binary stages, where each stage is preceded by a gate circuit removing one input pulse per cascade output pulse, the count may be arranged to have any integral value between 2^n and $2^{n+1} - 1$.

4. Practical Example

A complete counter, based upon the principles outlined, has been developed in connection with an investigation into the properties of dot-interlace[5] as a bandwidth-saving device. Such an application is very similar in requirements to those outlined in Section 1 although, in this case, the operating range of input-pulse repetition frequency is somewhat higher. A block schematic diagram of the counter (Fig. 5) shows a cascade of four binary stages arranged to count by any integer within the limits of 16 and 31 inclusive.

The input-pulse train is applied to the first gate circuit G_1 wherein one input pulse may be removed for each gating pulse derived from P_1; that is, for each cascade output pulse. The output of the gate G_1 drives the first binary stage S_1.

Short pulses representing the output of S_1 now form the input to the double-throw switch circuit

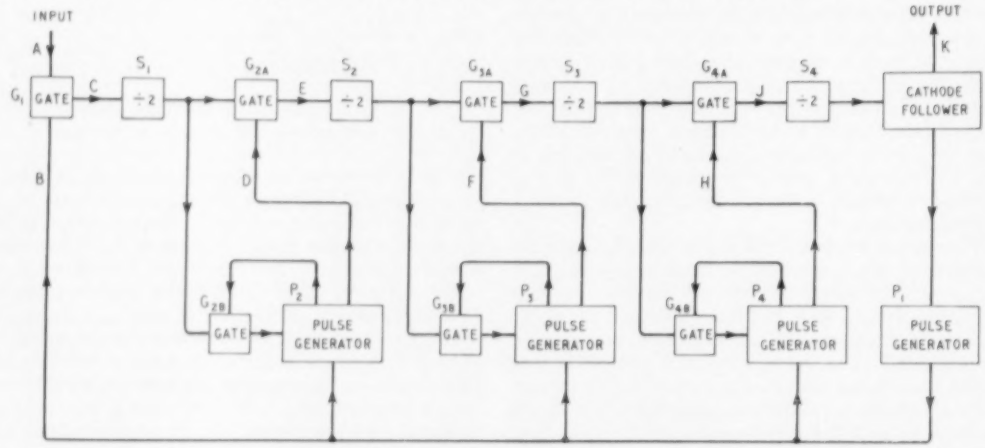

Fig. 5. *Schematic diagram of a gated counter.*

formed by G_{2A} and G_{2B}. The principle of operation of this arrangement has been outlined above and illustrated in Fig. 4. The two gates G_{2A} and G_{2B} have identical input waveforms but are operated by control pulses of opposite polarity such that, in the normal condition, G_{2A} is open and G_{2B} is closed. A cascade output pulse fed through P_1 will actuate P_2 causing G_{2A} to close and G_{2B} to open. The next input pulse to the gates will now be routed to P_2 thus terminating the control pulse. The circuit has now been returned to the normal condition (G_{2A} open, G_{2B} closed) and will remain so until the next cascade output pulse.

The remaining binary and gate stages of the cascade are similar in design and performance. By means of switches each of the gating circuits may be arranged either to remain open continuously or to operate as described. Thus, one stage input pulse per cascade output pulse may be removed at will from the input to any of the binary stages in the cascade. In this way the cascade count may be varied over the range of 16 (all gates open continuously) to 31 (all gates pulse-operated) inclusive.

In order to furnish more detailed design information, circuits of the basic binary stage and the two forms of gating arrangement are described below.

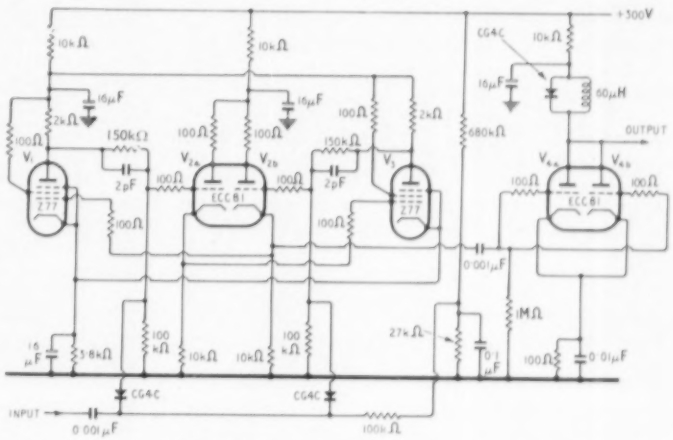

Fig. 6. *Binary stage suitable for high-frequency operation.*

4.1. *Binary Stage*

Fig. 6 shows the circuit of a binary stage as incorporated in the cascade. The valves V_1, V_2 and V_3 form a directly-coupled multivibrator wherein the cross-coupling feedback paths are routed through the cathode followers formed by V_{2A} and V_{2B}. By this means, the recovery time of the circuit is reduced to a minimum in that grid current in either V_2 or V_3 cannot cause charging of the cross-coupling capacitors. In addition, the capacitance appearing across the load resistors of V_1 and V_3 is reduced to a minimum. The input pulses, of negative polarity, are applied through crystal rectifiers to the grids of the cathode followers V_{2A} and V_{2B}. The bias on the rectifiers is such that a negative input pulse is applied to that grid of V_2 having the greater positive potential with respect to earth. The rectangular voltage waveform appearing at the cathode of V_{2B} forms the drive to V_4. The resulting anode current of V_4 flows through the pulse-generating circuit formed by the resonance of the anode load inductance with stray capacitance. A crystal rectifier is connected in such a way as to damp all oscillation other than that resulting from a cessation of anode current in V_4. A positive voltage pulse of half-sinusoid shape appears at the anode of V_4 for every two pulses applied at the stage input.

4.2. *Gating Circuits*

As will be evident from Fig. 5, two types of gating arrangement are utilized in the complete counter. The first, used in the gate G_1, is of the type discussed with reference to Fig. 3 and consists simply of an amplifier accepting input trigger pulses of positive sign at the control grid and control pulses of negative polarity at the suppressor grid. A control switch is provided whereby the control pulses may be removed from the suppresssor grid, resulting in a continuous and uninterrupted pulse output from the gate. Control pulses of positive polarity are derived from the pulse generator P_1 (Fig. 5) wherein the rectangular waveform at the cascade output is converted into a train of pulses having a duration, polarity and timing such that the gate G_1 removes one cascade input pulse for each cascade output pulse. The circuit arrangement used for P_1 is very similar to that associated with V_4 in Fig. 6. The output pulses from P_1 are also used as trigger pulses for the other gates (G_2, G_3 and G_4) in the counter. The circuit arrangement used for these gates is of the type illustrated in Fig. 4 and a detailed circuit diagram of one such stage is shown in Fig. 7.

Input pulses of positive polarity are applied to the control grid of V_1 and, in the absence of gating pulses at the suppressor grid, are passed from the anode circuit of V_1 to the next binary stage

(Fig. 4, gate switch A, position X). The ungated input pulses appear at the cathode of V_1 and are fed to an amplifier V_5. At this point in the discussion it will be assumed, for the sake of simplicity, that the high-speed bistable stage formed by V_2, V_3 and V_4, is in that state of equilibrium where V_2 is cut off and V_4 conducting. A trigger pulse derived from the cascade output is amplified in V_6 and is passed through a suitably-biased crystal rectifier to the anode of V_2, causing a change in the circuit conditions of V_2, V_3 and V_4 such that V_2 conducts and V_4 is cut off. The resulting negative-going voltage transition occurring at the cathode of V_{3A} is fed to the suppressor grid of V_1, cutting off the pulse drive to the following binary stage. A similar transition of positive polarity appearing at the cathode of V_{3B} is fed to the suppressor grid of V_5 so that the next input pulse appears at the anode of V_5 (Fig. 4, gate switch A, position Y). This pulse, fed through a suitably-biased rectifier, causes the bistable circuit to revert to the first condition of equilibrium. The resulting positive-going voltage transition at the cathode of V_{3A} terminates the gating pulse to V_1 and a similar negative transition at the cathode of V_{3B} results in anode current cut-off in V_5. The circuit has thus completed one gating cycle and will repeat the above operations whenever a trigger pulse is received from the cascade output. It will be appreciated that the circuit shown relies on rectification for bias setting at the grids of the coincidence gates V_1 and V_5. The rectification is effected either by grid current or by means of anode current in auxiliary diodes. The grid time-constants shown are suitable for operation with a cascade whose output repetition frequency is fairly high and these constants should be increased when the gating circuit is required to operate at lower repetition frequencies.

As in the case of the gate G_1, all the remaining gate circuits are provided with a control switch whereby the gating process may be discontinued. Such a switch is shown in Fig. 7, in series with the h.t. supply to valves V_2 and V_4. When this switch is opened, gating pulses are removed from the suppressor grid of V_1.

5. Results and Conclusions

The complete counter unit, as outlined above and illustrated schematically in Fig. 5, has proved to be fully practicable and reliable. In order to provide some evidence of correct operation and to illustrate further the operating sequence, a series of oscilloscope photographs is provided in Fig. 8. The traces show waveforms at various points in the complete circuit when an input-pulse train having a repetition frequency of 5·5 Mc/s is applied to the counter, a value of input repetition frequency equal to the video sampling rate used in some experiments carried out on a dot interlace television system as mentioned in Section 4.

In the waveforms illustrated all the gate circuits are in the pulse-operated condition and, as there are four binary stages ($n = 4$), the overall divisor has the value 31 ($2^{n+1} - 1$). Thus, for an input repetition frequency of 5·5 Mc/s, the repetition frequency at the output will be 177 kc/s.

The code letters beside each photograph refer to a similar set in Fig. 5 and illustrate the waveforms appearing at the corresponding points indicated in that schematic diagram.

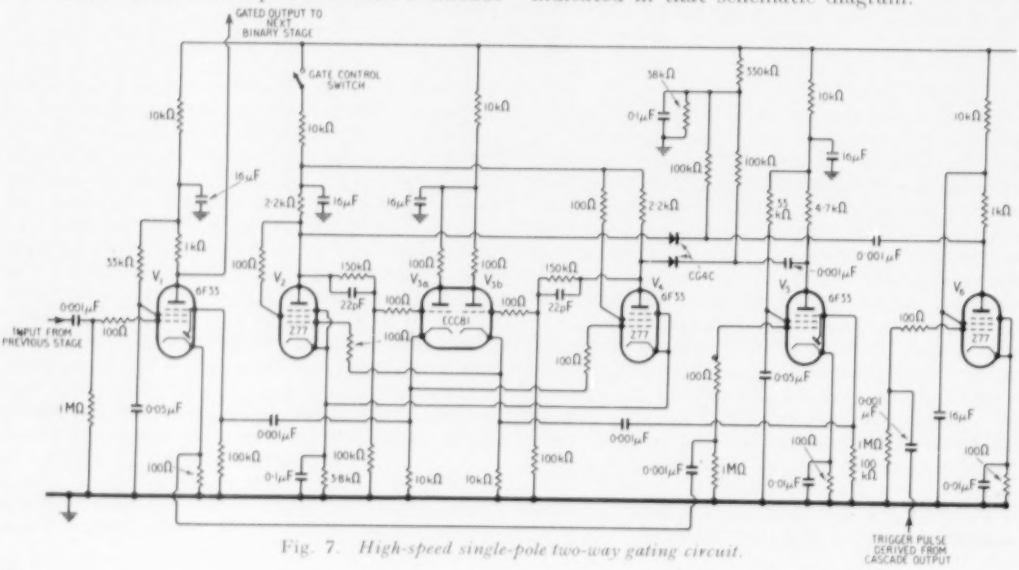

Fig. 7. *High-speed single-pole two-way gating circuit.*

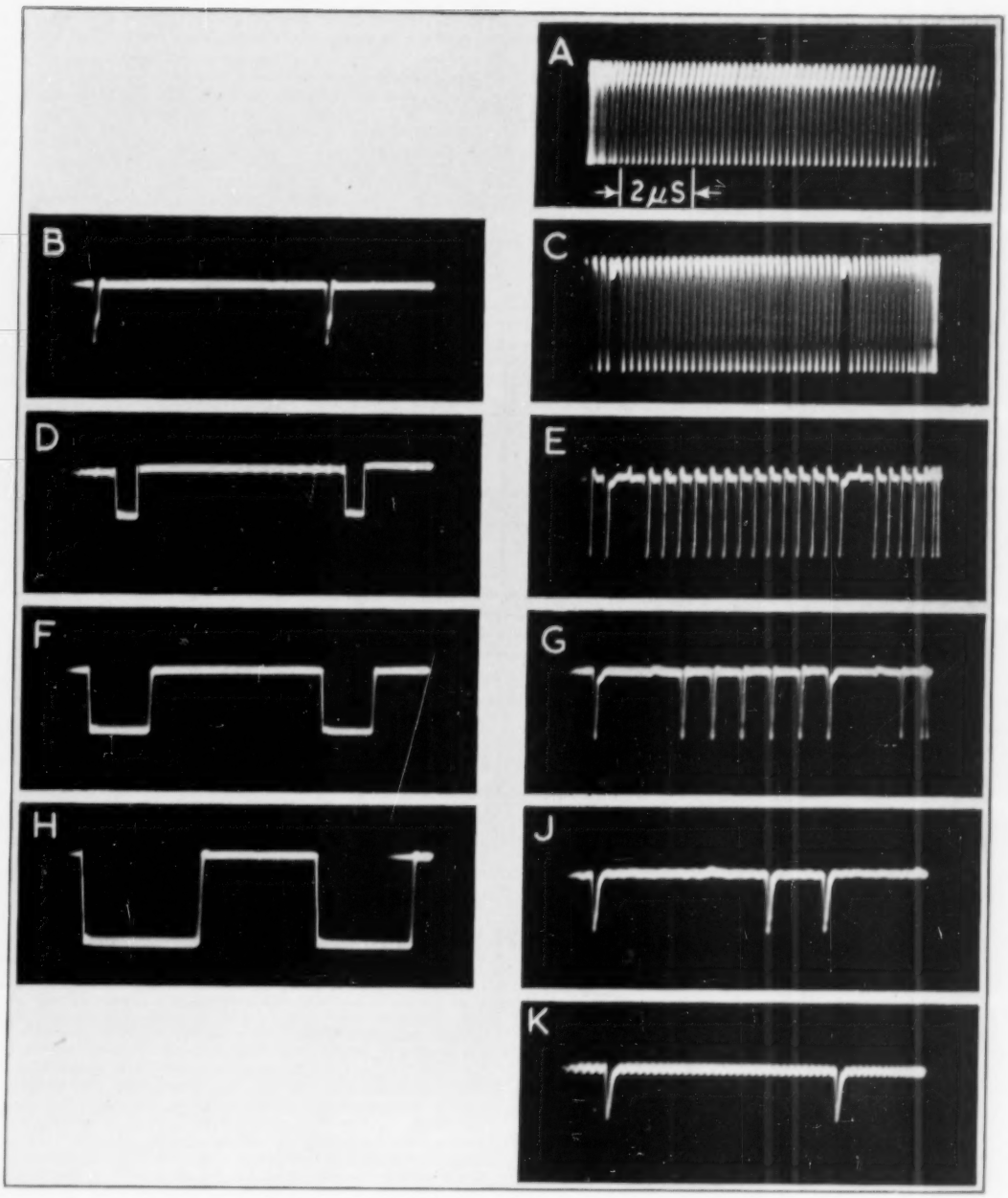

Fig. 8. *Waveform at various points within the counter of Fig. 5.*

A counting arrangement of the type described should prove valuable in many applications, particularly where high operating frequencies are encountered. The circuit provides convenient means for changing the overall count, a feature which may prove valuable in future colour-television systems. Further developments of the arrangement can effect a saving in the total number of valves employed.

A complete counter based on the foregoing work has been built and has given trouble free operation since its inception, a period of 700 working hours.

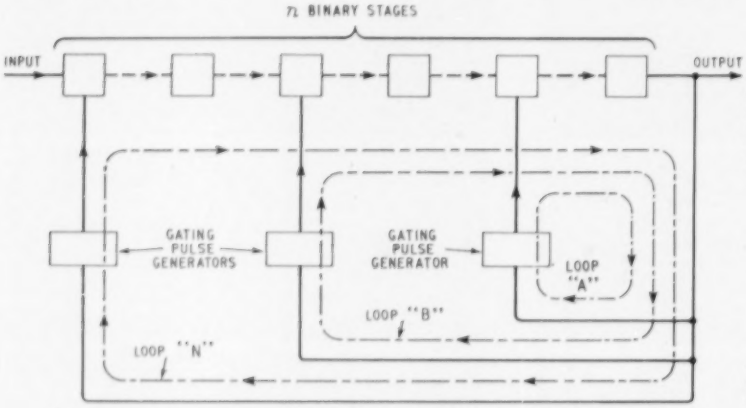

Fig. 9. *General arrangement of gating binary stages in a gated counter.*

Acknowledgment

The authors wish to thank the Chief Engineer of the British Broadcasting Corporation for permission to publish this article.

APPENDIX

Fig. 9 shows an arrangement of n binary stages, a gating circuit being provided at the input of certain stages.

For the purpose of analysis, loops A, B and N are indicated in the diagram, loop A embracing a stages, loop B embracing b stages, etc.

If each cascade output pulse suppresses one input pulse at the first gate in each loop, the divisor D_A of loop A will be

$$D_A = 2^a + 1 \qquad (1)$$

For loop B

$$D_B = 2^{b-a}(2^a + 1) + 1 = 2^b + 2^{b-a} + 1$$

Similarly for loop N

$$D_N = 2^n + 2^{n-a} + 2^{n-b} + 2^{n-c} + \ldots\ldots + 2^{n-m} + 1$$

where m is the number of binary stages enclosed by the penultimate loop. In general

$$D_N = 1 + 2^n + \sum_{\lambda=m}^{a} 2^{n-\lambda} \qquad (2)$$

When each stage of the cascade is preceded by a gating circuit removing one stage-input pulse per cascade-output pulse, the total number of loops will equal the total number of stages n.

Hence, from (2)

$$D_N = 1 + 2^n + \sum_{\lambda=n-1}^{1} 2^{n-\lambda} = \sum_{\lambda=0}^{n} 2^\lambda = 2^{n+1} - 1$$

since D_n is a geometrical progression of $n + 1$ terms, with first term 1, and ratio 2.

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CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

The Bifilar-T Trap

SIR,—Your readers may be interested in further information regarding the bifilar-T trap which you discussed in recent issues of *Wireless Engineer* and which is being used in colour television sets in this country: the bifilar-T trap, as used in television circuitry, is discussed in the following R.C.A. Laboratory Bulletins:—

Benjamin Fisher and Jack Avins, "An Analysis of the Bifilar-T Trap Circuit", RCA Industry Service Laboratory Bulletin LB-961, September 16, 1954.

Morris D. Nelson and Jack Avins, "The Design of IF Amplifiers for Color Television Receivers", RCA Industry Service Laboratory Bulletin LB-950, May 20, 1954.

It may be of interest to know that the bifilar-T trap has been used at frequencies as low as 50 kc/s in this country. A composite filter which is used in the Hallicrafters' HT-30 transmitter in combination with other circuit elements for the generation of single-sideband suppressed-carrier telephony signals employs the 50-kc/s version of the bifilar-T trap. The filter, which was designed by Wm. Counts of the Hallicrafters Company, achieves over 40-dB suppression of the unwanted sideband and has a cut-off of approximately 50 dB in 300 c/s, starting from the 3-dB point on the carrier side of the passband.

RICHARD F. BURNS

Beloit Research and Development Co.,
Beloit, Wisconsin, U.S.A.

24th July 1956.

Constant-Frequency Oscillators

SIR,—Mr. Lukaszewicz's assumption, in the August issue of *Wireless Engineer*, p. 201, that the current ratio n_{g1} is real is equivalent to assuming from the outset that the oscillation frequency is independent of μ: for the expression for i_r shows, when the harmonic voltages are zero, that i_{a1} $(1 + n_{g1})$, and so also i_{a1} and i_{g1}, must have the same phase as $v_{g1} + v_{a1}/\mu$. As neither i_a nor i_g are functions of $v_g + v_a/\mu$ this requirement can be satisfied only if both v_{g1} and v_{a1} are in phase (or antiphase) with i_{a1}. Changes in μ can then affect only the magnitudes but not the phases of i_{a1} and i_{g1}, and so the oscillation frequency is independent of μ. Thus Mr. Lukaszewicz's proof is based on circular reasoning.

With regard to the effects of harmonics, none of the papers to which Mr. Lukaszewicz refers shows how the oscillation frequency can be made independent of the harmonic content of the input and output voltages. However, the criterion derived for the stability of sinusoidal oscillations can easily be extended to cover nonsinusoidal operation.

Harmonic voltages influence the oscillation frequency by changing the phases of the input and output currents of fundamental frequency through intermodulation in the nonlinear amplifier. If the input and output voltages are of the form $V_1 \cos \omega t + \sum_2^\infty V_n \cos (n\omega t + \theta_n)$ the condition for the input and output currents of fundamental frequency to be of the form $I_1 \cos \omega t$, independent of the values of V_n, is that $\theta_n = 0$ for all n. Thus the harmonic currents and voltages must have the same phase, from which it follows that the input, output and transfer impedances of the network must be resistive at the oscillation frequency and all its harmonics.

A. S. GLADWIN

University of Sheffield.
23rd July 1956.

NEW BOOKS

Taschenbuch der Hochfrequenztechnik

By H. MEINKE and F. W. GUNDLACH. Pp. 1408 + xxviii with 1856 illustrations. Springer-Verlag, Reichpietschufer 20, Berlin, W.35. Price D.M. 69.

Although called a 'pocket-book' it should be noted that it has 1436 pages and weighs 3 lb. Professors Meinke and Gundlach are the directors of the high-frequency institutes of the technical universities of Munich and Berlin-Charlottenburg respectively, but the book is the combined output of thirty-seven German experts in the various fields, of whom particulars are given in the introduction. The book is not to be regarded as a text-book but as a reference book in which the high-frequency engineer can find curves, formulae and data which will give him the scientific basis for the research or development work on which he is engaged. It might justifiably be called an encyclopaedia in one volume.

The book is divided into 25 chapters, not numbered but lettered from A to Z, and each chapter is then divided into numbered sections. These are set out in a 17-page table of contents, giving also the names of the authors who have contributed to each chapter: the subdivision of the chapter between the various authors is given in a footnote at the beginning of each chapter. Each chapter ends with an extensive index to the literature referred to, the total number of references being over 2,000. The book concludes with the usual subject index. A slip giving a list of nine errata is inserted in the book, but it overlooks the two spellings, Oersted and Oerstedt within four pages of one another, doubtless due to different authors: there is also a mistake in the chapter heading of p. 1312. On the whole, the book appears to have been prepared with great care.

The opening chapter entitled "circuit elements" begins with the fundamental laws of magnetism and skin effect, and then, after 42 pages devoted to coils, etc., of various types, devotes the next 40 pages to the electric field and capacitors: then follows h.f. resistance, etc., piezo effect, and quartz crystals, and the chapter concludes with 119 references. Chap. B deals with circuits with concentrated or lumped elements, whereas Chap. C deals with homogeneous conductors, especially h.f. cables: in both cases the treatment is very thorough both mathematically and graphically. The following four chapters are devoted to waveguides, both hollow and solid of the Goubau type, the various elements and devices that enter into the construction of lines and waveguides and with cavity resonators and filters. Chap. H of 143 pages deals very thoroughly with every type of h.f. aerial, and the following chapter with the radiated waves, their reflection, refraction, etc.

With Chap. K we turn to semiconductors, thermistors, transistors, etc., followed by the construction and properties of ordinary valves, and then those of the klystron and allied types; then follow chapters on amplifiers of all types and applications, and on the use of valves as rectifiers. Chap. Q deals with the superposition or mixing of different frequencies. Chap. R with the production of oscillatory currents, Chap. S with impulse circuits such as the flip-flop and multivibrator, and Chap. T with the various sources of noise. Chap. U of 100 pages is devoted to modulation and demodulation, while Chap. V is a relatively short chapter by Kupfmüller on the general principles of information transmission. Chapters W and X are also short chapters entitled "senders" and "receivers" respectively, discussing such things as the various types of modulation, long, medium, and short waves, single-sideband transmission and reception. Chap. Y of about 100 pages by 7 contributors deals with h.f. measurement of every kind. The final chapter is a very thorough discussion of transformers and rectifiers with choking coils, smoothing capacitors and filters.

The book is an outstanding production, containing the essential material of every branch of the subject, arranged and set out in an exemplary manner, together with a liberal supply of international references.

G.W.O.H.

Television Engineering: Vol. 2. Video-Frequency Amplification

By S. W. AMOS, B.Sc.(Hons.), A.M.I.E.E. and D. C. BIRKINSHAW, M.B.E., M.A., M.I.E.E. A B.B.C. Engineering Training Manual. Pp. 270. Published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 35s.

Das Ohr als Nachrichtenempfänger

By R. FELDTKELLER and E. ZWICKER. Pp. 90. S. Hirzel Verlag, Birkenwaldstrasse 185, Stuttgart N., Germany. Price D.M.14.

Wireless Servicing Manual (9th Edn.)

By W. T. COCKING, M.I.E.E. Pp. 268 + viii. Published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 17s. 6d.

R.F. Transmission Lines
Edited by ALEXANDER SCHURE, Ph.D., Ed.D. Electronic Technology Series No. 166–8. Pp. 63. Price $1.25.

Amplitude Modulation
Edited by ALEXANDER SCHURE, Ph.D., Ed.D. Electronic Technology Series No. 166–9. Pp. 56. Price $1.25.

Blocking Oscillators
Edited by ALEXANDER SCHURE, Ph.D., Ed.D. Electronic Technology Series No. 166–10. Pp. 64. Price $1.25.

Picture Book of TV Troubles: Vol. 5, Horizontal Output and H-V Circuits
By John F. Rider Laboratories Staff. Pp. 108. Price $1.80.

Radio Receiver Laboratory Manual
By ALEX. W. LEVEY, B.A., M.S. Pp. 112. Price $2.
The above five books can be obtained from John F. Rider Publisher Inc., 480 Canal Street, New York 13, N.Y., U.S.A.

Guide to Broadcasting Stations 1956–1957. 9th Edition
Compiled by the staff of *Wireless World*. Pp. 80. Published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 2s. 6d. (postage 4d.).

British Plastics Year Book 1956. 26th Edition
Pp. 740. Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 35s.

Electrical Who's Who 1956-57
Pp. 458. Published by *Electrical Review* Publications Ltd., and distributed by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 21s.

Metal Transfer between Palladium and Silver Contacts at Low Inductance
By J. RIDDLESTONE. Report No. U/T133. Pp. 10. Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 12s. 6d.

A Stable Synchronous Detector for Audio-Frequency Measurements
By P. G. KENDALL. Report No. V/T125. Pp. 18. Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 12s. 6d.

Radio Valve Data. 5th Edition
Compiled by the staff of *Wireless World*, this book contains full operating data on over 2,500 types of British and American radio valves, 37 transistors and 300 cathode-ray tubes. The main tables give electrical characteristics of each valve and classify the valves into current, replacement or obsolete types, as recommended by the makers. An index enables any valve to be found immediately in the tables, while the full list of equivalents has been enlarged. Base connections are shown diagrammatically, the main tables being keyed to the series of diagrams.
Published for *Wireless World* by Iliffe & Sons Ltd., Dorset House, Stamford Street, London, S.E.1. Price 4s. 6d. (postage 6d.).

OBITUARY

We regret to record the death of Labouchere Hillyer Bainbridge-Bell, O.B.E., M.C., M.A., on 7th August after a short illness. Born in 1893, "B-B" served in World War I with R.E. Signals and later as wireless officer in the R.F.C. and R.A.F. He was on the research staff of the Radio Communication Co. from 1920–26 and then joined the Radio Research Station, Slough.
He was one of the pioneer workers on radar from 1935–1939, after which he joined the Admiralty Signals and Radar Establishment, where he remained until his retirement in 1953. The Royal Commission on Awards to Inventors awarded him £2,400 for his contributions to the development of radar transmitters and receivers, and an optical convertor.
Since his "retirement" he has been a committee secretary in the electrical division of the British Standards Institution.

MEETINGS

Brit.I.R.E.

26th September. "Some Aspects of Transistor Progress," by H. W. Lceb, Ph.D., to be held at 6.30 at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

Television Society

21st September. "Recent Advances in Colour Display Tubes", by Dr. R. R. Law, at 7 o'clock at 164 Shaftesbury Avenue, London, W.C.2.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for July 1956

Date 1956 July	MSF 60 kc/s Frequency deviation from nominal*: parts in 10^9
1	N.M.
2	−3
3	−2
4	−2
5	−2
6	−1
7	N.M.
8	N.M.
9	−2
10	−2
11	−2
12	−2
13	−2
14	−1
15	−1
16	−2
17	−2
18	−2
19	−2
20	−2
21	−1
22	−1
23	−1
24	−1
25	−2
26	−2
27	−2
28	N.M.
29	−2
30	−1
31	−1

N.M. = Not Measured.

*Nominal frequency is defined to be that frequency corresponding to a value of 9 192 631 830 c/s for the N.P.L. caesium resonator.

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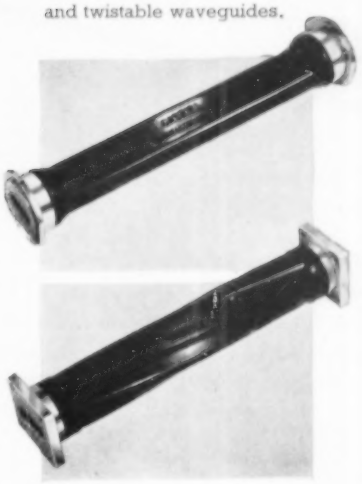

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  Flexible – non-twistable.
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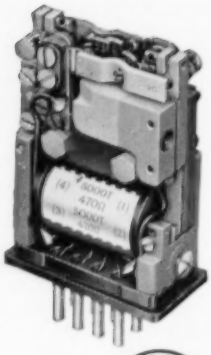

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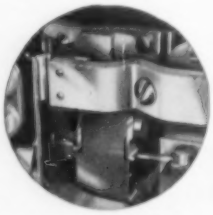

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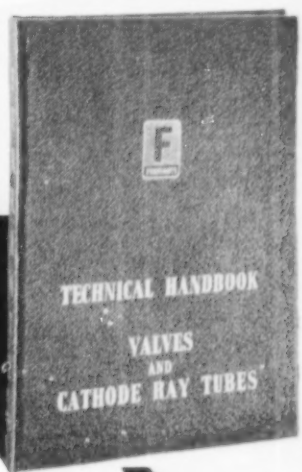

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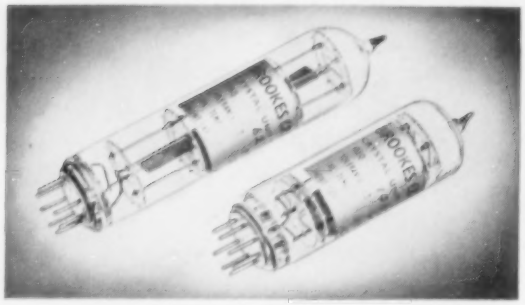

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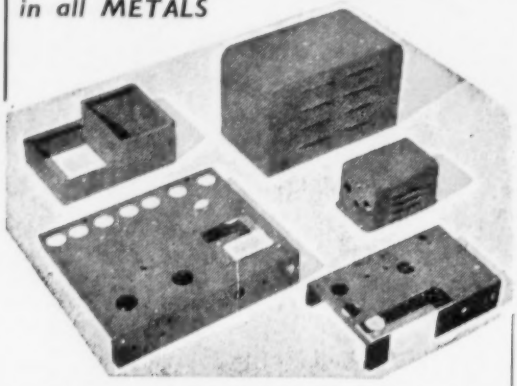

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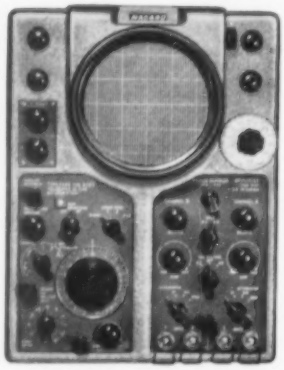

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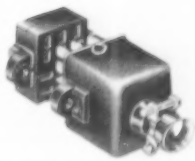

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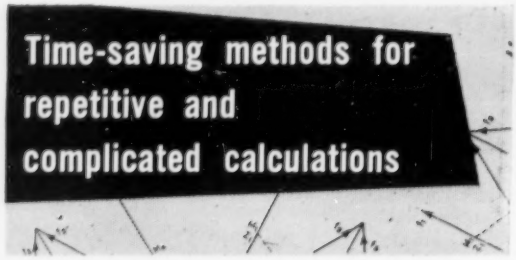

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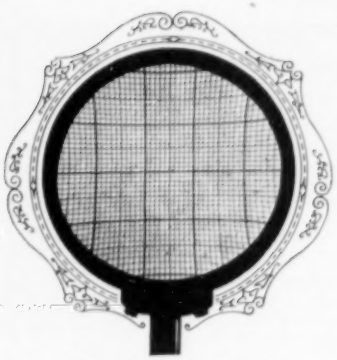

Actual camera shot of graph paper through a spherical magnifier at 2X magnification.

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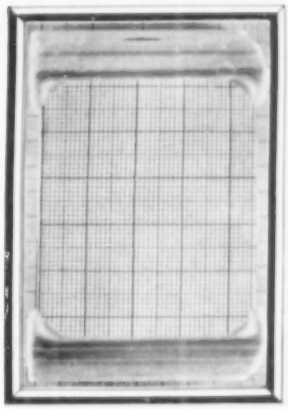
through

the

"Looking Glass"...

With apologies to Lewis Carroll



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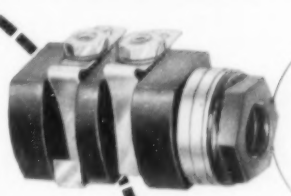

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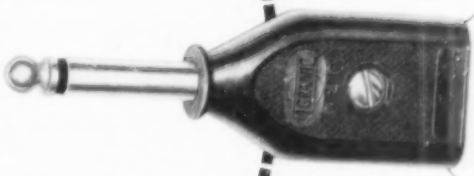

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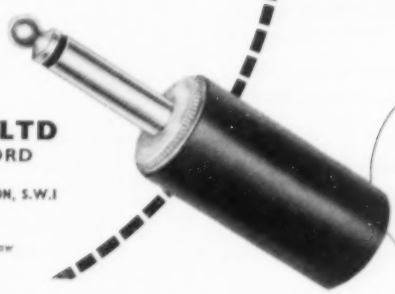

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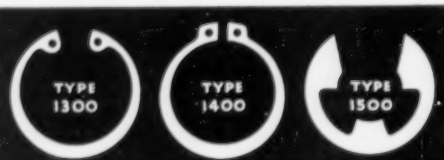

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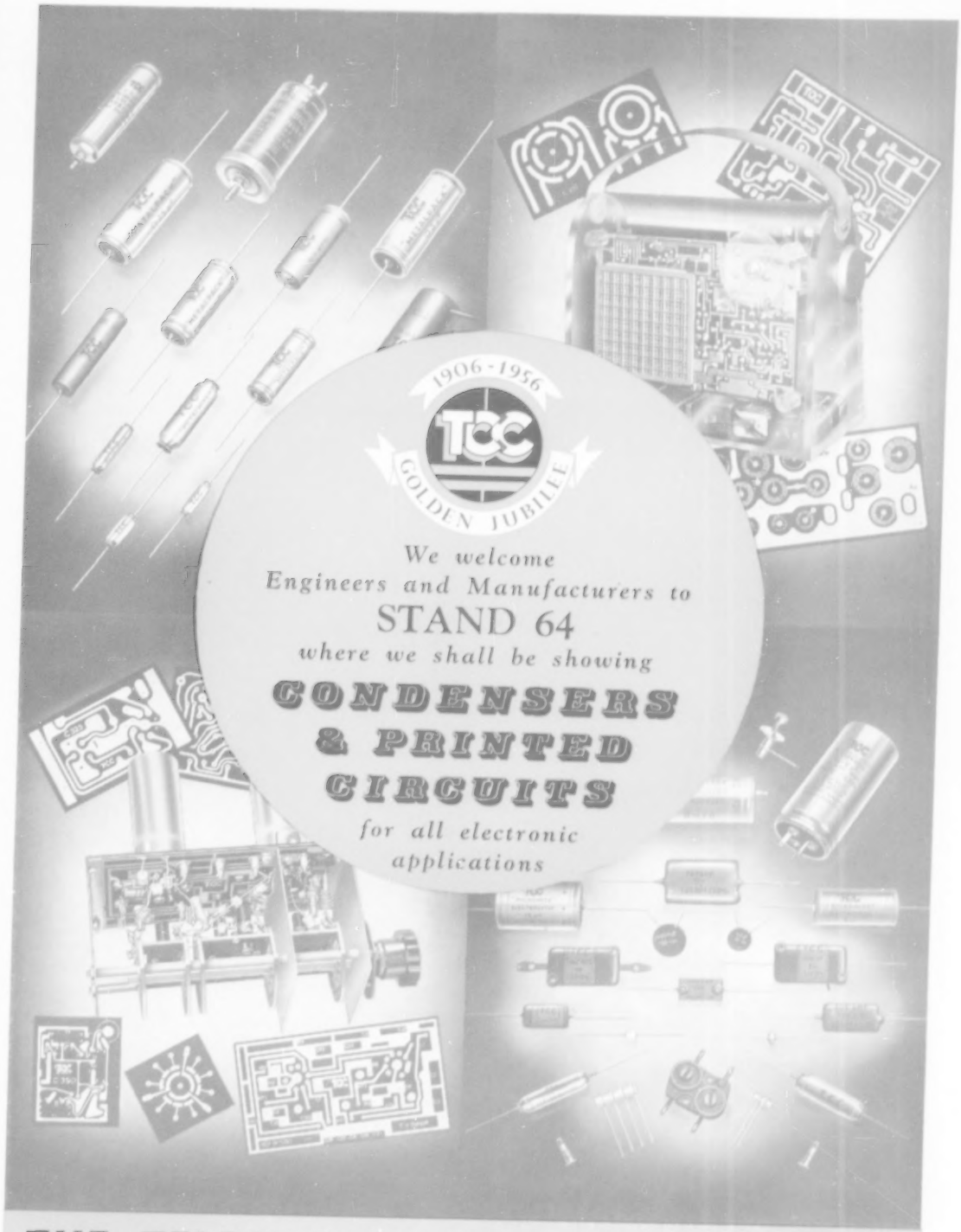

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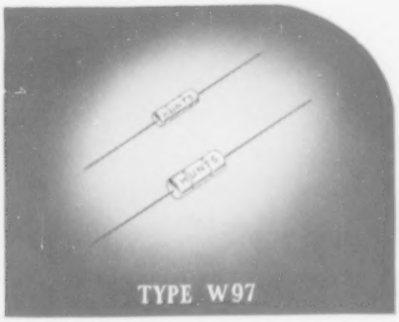

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BM8	0.004	0.610	0.135
BM11	0.004	0.500	0.180
BM9	0.005	0.610	0.135
BM12	0.005	0.500	0.180
BM13	0.01	0.500	0.180
BM14	0.02	0.610	0.180
BM15	0.03	0.610	0.260
BM16	0.04	0.610	0.260
	400 volts D.C.	Wkg.	
BM4	0.0004	0.610	0.135
BM5	0.0005	0.610	0.135
BM6	0.001	0.610	0.135
BM18	0.002	0.500	0.180
BM19	0.003	0.500	0.180
BM20	0.005	0.610	0.180
BM21	0.01	0.610	0.260
	600 volts D.C.	Wkg.	
BM25	50 pF.	0.500	0.180
BM1	0.0001	0.610	0.135
BM26	0.0001	0.500	0.180
BM2	0.0002	0.610	0.135
BM27	0.0002	0.500	0.180
BM28	0.00022	0.500	0.180
BM29	0.00025	0.500	0.180
BM3	0.0003	0.610	0.135
BM30	0.0003	0.500	0.180
BM36	0.0004	0.500	0.180
BM31	0.0005	0.500	0.180
BM32	0.001	0.500	0.180
BM33	0.002	0.610	0.260
BM34	0.003	0.610	0.260
BM35	0.004	0.610	0.260

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